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Monterey, California



THESIS

K14933

JOINT TACTICAL ANTI-JAM COMMUNICATIONS:
A SYSTEMS APPROACH

by

Vasilios Christos Kavvadias

September 1989

Thesis Co-Advisor: Donald v.Z. Wadsworth
Thesis Co-Advisor: William Gates

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This study analyzes the conceptual design of a joint tactical anti-jam communication link from a systems point of view. It addresses the requirements and the specifications for a communication system providing an integrated solution for navies operating in closed-sea areas under intense enemy jamming activity. The concept of the proposed system is based on spread spectrum technology and on the Joint Tactical Information Distribution System (JTIDS). Spread spectrum technology has been an area of extensive research for many years. Satisfactory practical solutions have been provided through the implementation of several frequency hopping systems that give partial answer to the anti-jam (AJ) problem. JTIDS is the only hybrid spread spectrum system intended to provide a catholic answer. The AJ performance of the proposed system is examined theoretically under realistic scenarios. System feasibility, from the overall cost standpoint, is evaluated using life cycle costing and sensitivity analysis. The trade-off between the procurement of an original system and a JTIDS-based design is also evaluated, based on the possible research costs. It is assumed that acquisition or procurement of such a system is not limited by any technology transfer barriers.				
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Joint Tactical Anti-jam Communications:
A Systems Approach

by

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Lieutenant, Hellenic Navy
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Submitted in partial fulfillment of the
requirements for the degree of

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September 1989

ABSTRACT

This study analyzes the conceptual design of a joint tactical anti-jam communication link from a systems point of view. It addresses the requirements and the specifications for a communication system providing an integrated solution for navies operating in closed-sea areas under intense enemy jamming activity.

The concept of the proposed system is based on spread spectrum technology and on the Joint Tactical Information Distribution System (JTIDS). Spread spectrum technology has been an area of extensive research for many years. Satisfactory practical solutions have been provided through the implementation of several frequency hopping systems that give partial answer to the anti-jam (AJ) problem. JTIDS is the only hybrid spread spectrum system intended to provide a catholic answer. The AJ performance of the proposed system is examined theoretically under realistic scenarios.

System feasibility, from the overall cost standpoint, is evaluated using life cycle costing and sensitivity analysis. The trade-off between the procurement of an original system and a JTIDS-based design is also evaluated, based on the possible research costs. It is assumed that acquisition or procurement of such a system is not limited by any technology transfer barriers.

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I. CONCEPTUAL MODEL DESIGN

A. INTRODUCTION

Tactical communications are essential to carrying out joint military operations. Furthermore, the tactical communication links must be secure as well as robust in the face of countermeasures. The establishment of anti-jam communication systems is particularly desirable in the case of navies which operate in closed sea areas under intense omni-directional jamming conditions where rapid, timely, secure, and survivable communications are required.

This thesis proposes a conceptual model design for an anti-jam communication system, based on the JTIDS (Joint Tactical Distribution System), for navies which operate in closed sea areas and environments of this nature. The operational considerations and the technical specifications in employing and deploying such a system are analyzed in this chapter. The theory of anti-jam (AJ) systems is presented in Chapters II and III. The analysis of the proposed system's engineering design and its subsequent testing under jamming conditions is described in Chapter IV. Chapter V is dedicated to the development of a model for a life cycle cost analysis of the system. The author's conclusions are presented in Chapter VI.

B. BACKGROUND

The need for tactical communications has been traditionally seen within the framework of each particular service of the Armed Forces. Even within the same service, differing needs and varying tasks have created a "mosaic" of equipment. This diversity led to the development and implementation of totally incompatible systems.

The lack of interoperability is primarily felt in joint operations where various communication platforms operate in a common area and serve a common task. Joint operations by their very nature imply complexity and task diversity. Since the operational requirements of the individual units, weapon systems and command levels have been traditionally met through diverse engineering

solutions, interoperability among them remained far from reality. Inefficient use of these resources is a natural result.

Only recently, navies have started looking at the need for communication within the context of Naval C3 (Command, Control and Communications) architecture. Command and Control by definition are concerned with those functions that aim to "the direction and control of general-purpose forces in accomplishment of the mission" [Ref. 1: p.3]. A commander is able to perform these functions in a timely fashion using communications. Communication serves as the backbone of the overall C3 architecture. A more technical examination of the elements of this architecture leads to an appreciation of the communication systems' vital role. A proper C3 architecture must be designed to solve the interoperability problems for all its three dimensions.

Generally speaking, a C3 system is a closed loop process which contains the following basic operations : Sense-Process-Classify-Evaluate-Plan-Decide and Act (S-P-C-E-P-D-A) [Ref. 1: p.4]. This generic process appears at every level of command. A wide range of sensors available to the commander provides the necessary environmental information. This is then processed through the system manually or automatically by ADP (Automatic Data Processing) stages, correlated with other events, filtered and finally displayed suitably to an evaluator. At the planning phase the information is multiplexed with existing doctrines and several alternative operational plans are developed. The decision making phase selects one alternative which becomes an operational order. This is communicated to the assigned assets and an action is taken. The effects of this action on the external environment are monitored and the results are fed back to the system for control purposes. This closes the loop. This process is shown in Figure 1. Typically the need for communication occurs at least twice in a single looping rotation: to gather the environmental information and to communicate decisions to the forces.

Whether an action is an actual application of a remote weapon system, a rapid redistribution of assets, or a tactical maneuver, all actions require timely and accurate C3 service. Establishing a C3 architecture requires developing

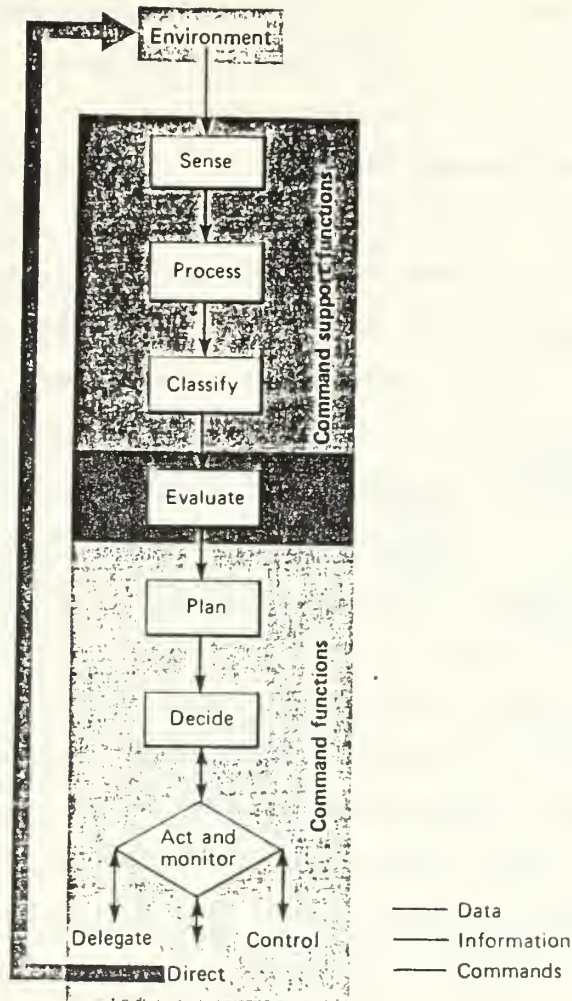


Figure 1. The Command and Control Process: A control system model showing the basic operations from sense through act [Ref. 1: p.4].

technologically feasible solutions for a survivable, secure, reliable, fast and interoperable C3 system. In the same manner, communication systems which serve as the ears, the eyes and the mouth of this integrated system must possess the features mentioned above.

C. NAVAL TACTICAL OPERATIONAL SCENARIO

Joint tactical operations in closed sea areas require timely, accurate and secure responses and actions. This task can be served by the availability and efficient integration of all the resources needed.

1. The C3 Fundamental Elements

Three distinct categories of elements enter into the C3 architecture. These are the naval command element, the naval forces element and the functional element connecting the previous two. In the joint tactical arena the command element includes the operational command centers at all levels, supporting the Fleet Commander-in-Chief down to the unified or the individual platform commander for a given area of operations. Generally, the mission is to direct the naval forces available to perform two main naval functions, sea control and power projection. The forces element includes all the resources available to the naval commander, such as surveillance units, shipborne weapon systems, ships, shore-based aircraft assigned to naval operations, and amphibious forces that support the mission. The generic C3 model can be thought in terms of a nodal network, as in Figure 2. A sequence of four nodes representing one shore based and three shipborne command centers is shown. These centers are the General Staff Command Center (GSCC), the Fleet Command Center (FCC), the Tactical Flag Command Center (TFCC) and the Combat Information Center (CIC). An integrated tactical communication system serves as the interface among these command centers which in turn support the operational commanders by providing information processing, storage, display and communication capabilities.

2. Description of the C3 Elements

A typical command center interfaces with various information sources through this communication linkage and provides the respective commanders with multi-source and fully correlated information, extending their information availability beyond the force's sensing limits. An integrated network of information collection, formed by over-the-horizon surveillance sensors and the intraforce sensors down to the individual platform-limited ones, provides the

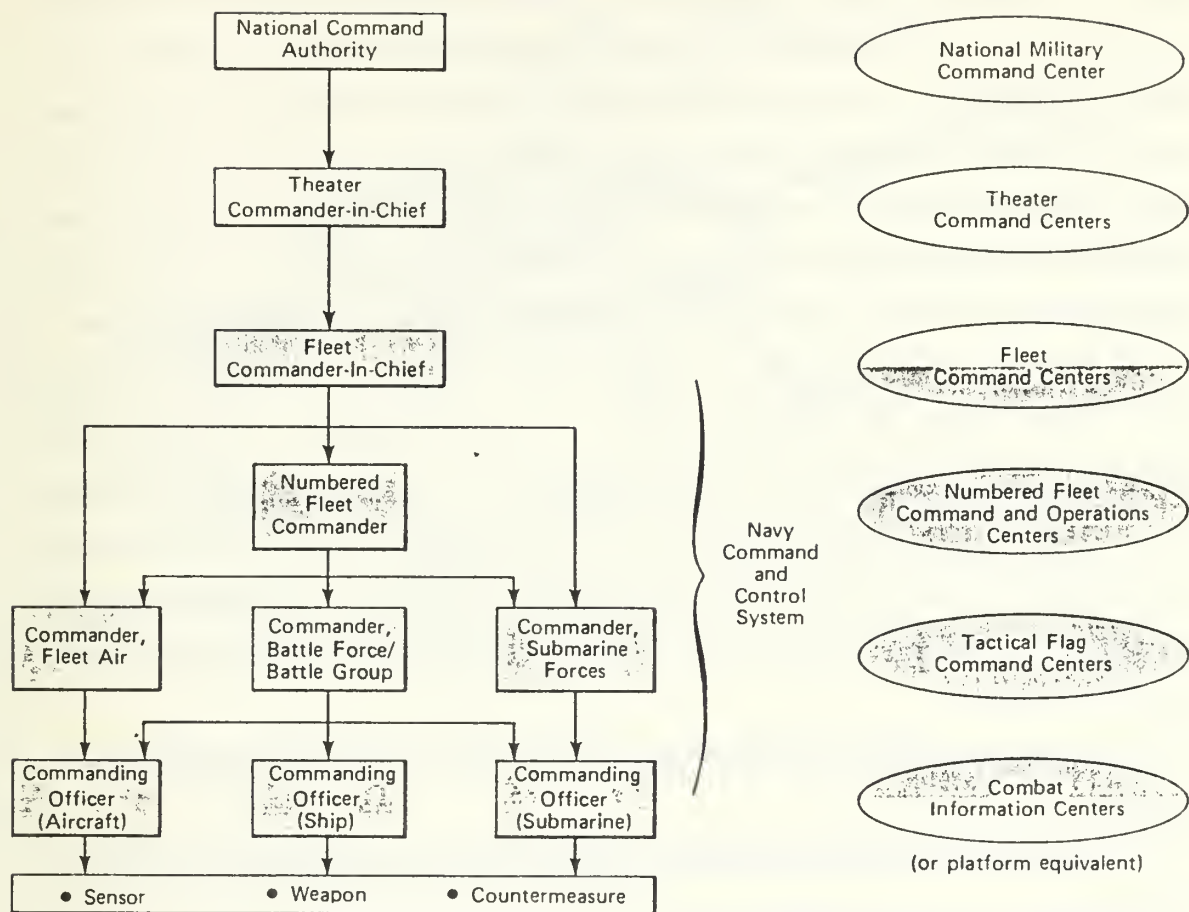


Figure 2. C3 Nodal Representation: A model showing the communications interface among the command centers.

overall tactical picture for every command level. This picture is the end product of a filtering process which actually occurs at the respective command centers.

The vital role of communications prevails when looking at the needs of the naval commander. The requirements for real-time two way communication among the intraforce and beyond the horizon sensor-platforms, and the constant monitoring of them within the context of a C3 system, should not be limited in any way. For instance, the commander-at-sea needs to be able to combine a sonar contact on a surface ship with a MAD (Magnetic Anomaly Detector) contact of a surveillance/reconnaissance MPA (Maritime Patrol Air-

craft), to identify a possible submarine position (POSSUB). The coordination of air, sea and land forces involved in joint naval operations is a rather complex task due to the interoperability issues resulting from the diverse policies, procedures and hardware used. Since policies are not firm, only the procedures and the hardware can be uniformly incorporated into the naval C3 architecture. This architecture then can essentially gather under its umbrella all the technological developments for the achievement of compatible solutions.

Since the naval assets generally operate beyond the horizon, long C3 paths exist. These may require relay communication nodes to provide integrated connectivity for all the forces assigned to a particular mission. The naval C3 system essentially serves all command levels from the joint tactical commanders down to the unit and platform commanders by enabling them to exercise their authority over the naval forces in the performance of their function [Ref. 2: p.2].

A typical operational scenario depicting joint naval operations is shown in Figure 3.

The next task is to specifically define where this scenario may apply and what the practical problems are.

D. PROBLEM DEFINITION

A typical geographical closed sea area in which the previously analyzed scenario applies conceptually, is shown in Figure 4. Several assumptions are made at this point:

- The C3 system's survivability is threatened by the existing omnidirectional jamming capabilities of the enemy forces.
- The enemy capability of signal interception and message exploitation is considered high.
- The friendly naval task force consists of a variety of air, sea and ground military forces and shore based C3 complexes, in a mixture of command, control, fighting and surveillance roles.
- Secure and fast responses of the joint task force can be challenged by the rapid deployment of enemy actions based on technologically advanced weapon systems.

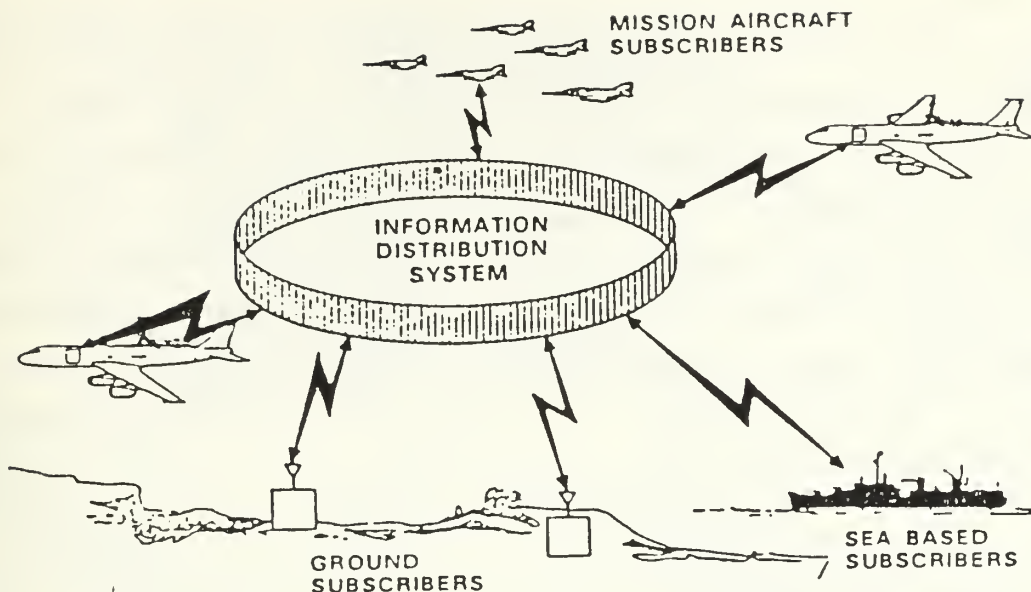


Figure 3. Typical Operational Scenario: A variety of air-sea and land assets involved in joint operations.

- Certain UHF spectrum availability exists for the accommodation of system's operation.
- Existing technology transfer barriers is not an issue since it is beyond the scope of this study.

What is required in this limited tactical arena is a joint tactical communication system which will serve as the backbone of the naval C3 architecture in the following manner:

- By providing an integrative linkage among the command centers and the forces involved, for effective coordination.
- By assisting the battle information management process through survivable, secure and real-time communications, for reliable operation execution and within realistic times before any effective enemy action.
- By establishing a platform-to-platform non nodal connectivity among the forces involved, for enhanced survivability in case that a central node is lost or damaged.
- By satisfying multiple link voice and data handling requirements.
- By providing such functions that can support the integration and common reference of precise tactical positioning and location of the assets, as well as identification for the avoidance of mutual interference.

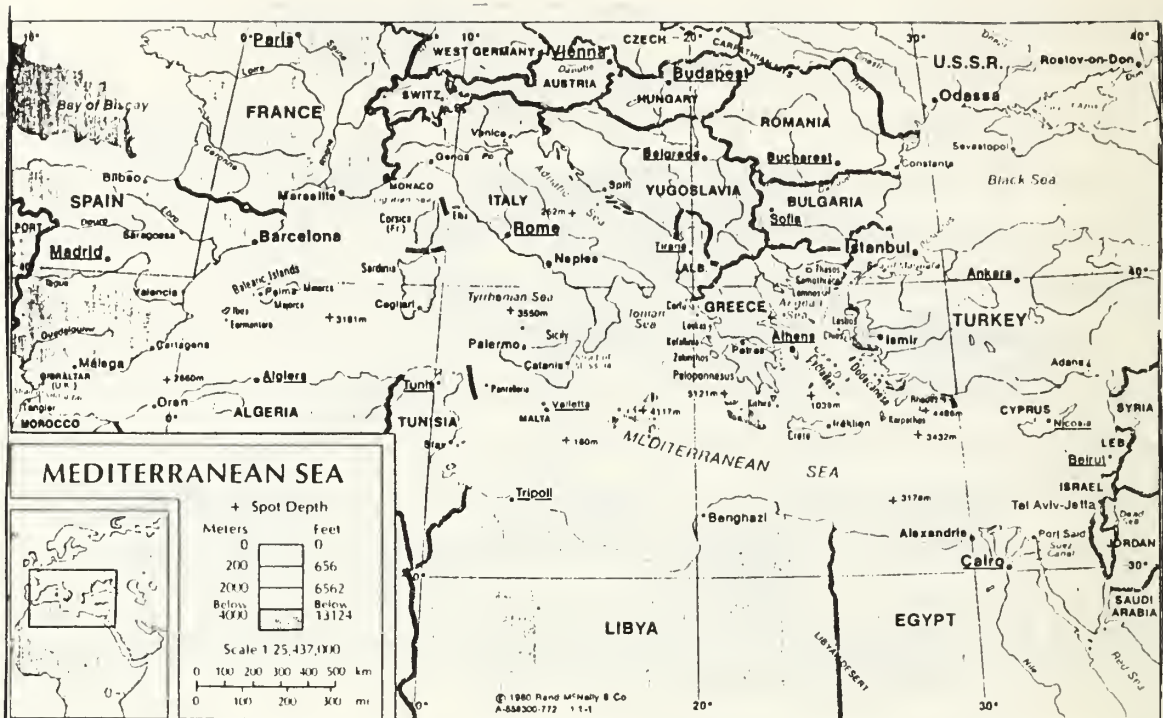


Figure 4. Example of a Closed Sea Area: A typical closed sea area of joint operations.

- By allowing for compatible connectivity with existing equipment used for the same purposes.
- By providing affordable solutions through the use of suitable technological options or possible alternatives, concerning procurement or leasing.

Before any solutions are proposed to this problem, employing sophisticated and advanced techniques and "state-of-the-art" technology, a more technical approach to the analysis of the naval C3 architecture is necessary. This will ease the translation from abstract operational considerations into more specific technical requirements.

E. ANALYSIS OF THE NAVAL C3 CONCEPTUAL ARCHITECTURE

1. Historical Factors

The rapid evolution of modern weaponry in the field of naval warfare constantly reduces the reaction times of a joint tactical naval force, especially

when it operates in limited geographical sea-areas. The need for timely decisions drives the development of an integrated C3 conceptual architecture.

Joint naval task forces are directed by those command centers coordinating a range of surveillance sources and weapon systems. The functional connectivity of these elements forms a conceptual C3 model. Several parameters enter into the C3 model and define its requirements. Naval doctrines, plans, policies and missions when coupled with the projected threat form the operational requirements [Ref. 3: p.9]. These in turn dictate the C3 architecture, which includes:

- The functions needed to achieve the stated objectives.
- The structure within which these functions are performed.
- The connectivities within the structure and the interfaces with the external entities for the command, control and information exchange.

2. Description of the C3 Process

The C3 closed loop process serves all command centers, by rotating the environmental information acquired by the surveillance assets and sensors. The C3 process as previously described filters the information through the S-P-C-E stages before the commander of a particular command center initiates plans, makes decisions and initiates the appropriate action. Although ADP is used extensively, human interface occurs at the evaluation phase and occasionally at the process and classification phases. Planning and decision making are eased by the use of existing doctrines. Ultimately the selected action is communicated through operational orders to the action performers. Halushynsky and Beam state that decisions can be implemented either by the delegation of authority to subordinate commanders, by the direction of a subordinate commander to carry out a selected action through an operational order, or by exercising direct control over subordinate assets. The loop closes by continuously sensing the results of the actions taken. Data exchange takes place at the "Process" and "Evaluate" stages. The evaluator correlates the information processed and received from both horizontally and vertically related command levels through a communication link. This link provides connectivity

among remote sensors, from the command center's own sensing stage and from several other command levels. The process is shown in Figure 5.

3. The C3 Functional Areas and Connectivities

Several functional areas enter into the C3 architecture. These are:
[Ref. 3: p.11]

- The Command area.
- The Information Management area.
- The Sensor Management area.
- The Engagement Management area.
- The Communications Management area.
- The System Management area.

The Command functional area includes planning, directing and assessing operations to achieve the assigned mission objectives.

The Information Management functional area includes acquiring, processing, and distributing data to ensure timely, accurate, and complete information, required by all users. It provides information collection, processing, evaluation and distribution services at each command level, resulting in an up-to-date tactical surveillance picture for the area of interest.

The Sensor Management functional area includes allocating, coordinating, and monitoring sensor assets to support decision making and weapons use. This requires that rules of engagement, emission control, and mutual interference with other sensors or communication assets are taken into consideration, before the surveillance data is fed into the Information Management area.

The Engagement Management functional area includes allocating, controlling, coordinating, and monitoring force assets to support execution of combat operations.

The Communications Management functional area includes allocating, controlling, coordinating, and monitoring the force assets such as decision aids, displays, and information handling systems, with the exception of communication assets.

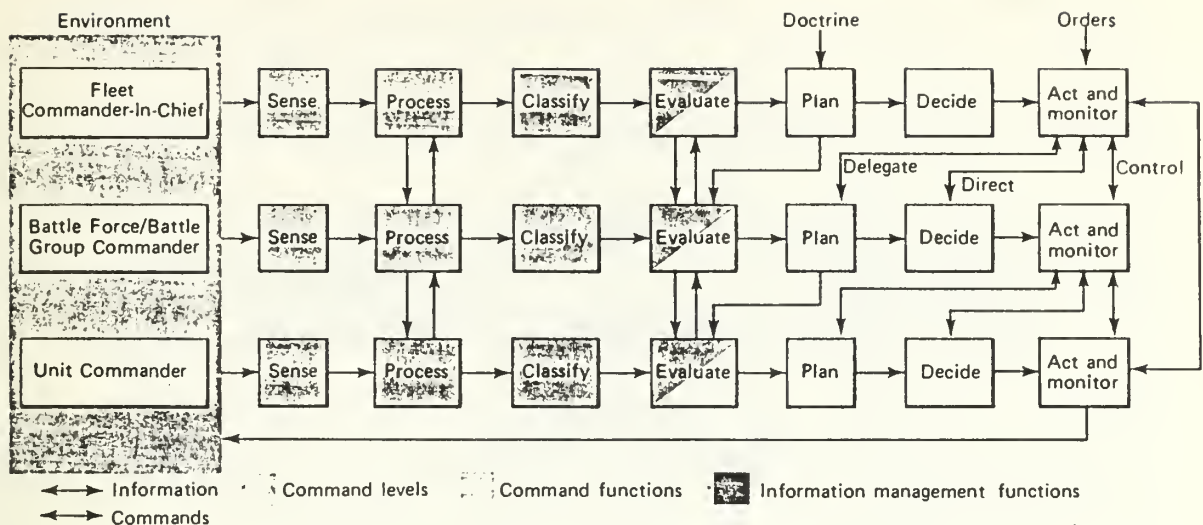


Figure 5. The Command and Control process: A functional Model [Ref. 3: p.11].

Finally, the System Management functional area allows a commander to assess the status of the C3 system state and reconfigure it in case of disruption.

The connectivities among the functional areas are shown in Figure 6. These are: [Ref 3: p.13]

- The connectivity among the Command functional areas.
- The connectivity among the Information Management functional areas.
- The connectivities among the Engagement, Sensor, Communications, and System Management functional areas at different command levels.
- The interfaces between command and the unified commands.
- The interfaces among Information sources and Information Management functional areas at various command levels.
- The interface between weapon systems and the Information and Engagement Management functional areas.

There are three types of connectivities:

- Command type - allowing for the direction and control of forces.

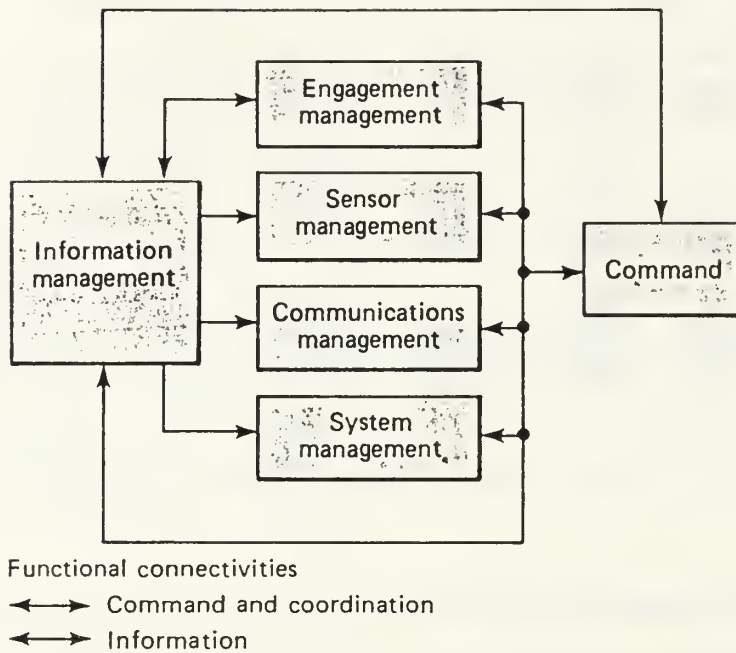


Figure 6. Functional Connectivities: Basic C3 structure [Ref. 3: p.12].

- Coordination type - between functional areas to ensure that assets are employed in accordance with command guidance.
- Information Exchange type - allowing for the transfer of data among the various areas to support the needs of the functional areas mentioned before.

4. The C3 Structure

Functional areas and connectivities can be expanded to all command levels forming the C3 structure. Generally speaking, the operational commanders directly involved in the joint naval tactical C3 in closed sea areas are:

- The Fleet Commander-in-Chief.
- The Battle Force/Battle Group (BF/BG) Commander.
- The Unit Commander.
- The Platform Commander.

According to Halushynsky and Beam, four options for connectivities apply to the naval C3 architecture of the future. [Ref. 3: p.14]

- Parallel connectivity between the Information Management areas at all command levels.
- Parallel downward dissemination for the interface between the information sources and the Information Management areas managed from their own levels.
- Hierarchical connectivity between adjacent asset management areas.
- Interface between weapon systems and the Information Management area at the platform level and between weapon systems and the Engagement Management areas at the battle force, battle group, and unit levels.

The information doesn't necessarily flow to all lower command levels. It is the responsibility of the appropriate commander to designate addressees and security levels. Furthermore, information must be geographically tailored to the capabilities of the recipient command level. Connectivity to commands below the BF/BG level from higher level sensors is maintained only when the surveillance units are tasked for the direct support of a particular unit or platform command.

The parallel connectivity of the Information Management areas at all command levels allows for timely information exchange. The connectivity of weapon systems to Information Management areas is limited to the level of the platform carrying the system. When the Officer in Tactical Command (OTC) directly controls certain weapon systems, especially those whose capability extends beyond the horizon or those which are dedicated to the defense of the BG, then a direct interface is maintained between the applicable weapon systems and the BF/BG command level or the unit level via the Engagement Management functional area.

5. The C3 Naval Architecture

The conceptual C3 naval architecture is depicted by mapping the actual naval operational commands and the various surveillance sources on the architecture previously analyzed. Part of this architecture is shown in Figure 7, covering from the Fleet Commander-in-Chief level down to the platform level, including both ground and afloat nodes.

Information, sensor and weapons management are essentially the main parts of this architecture. These are described below.

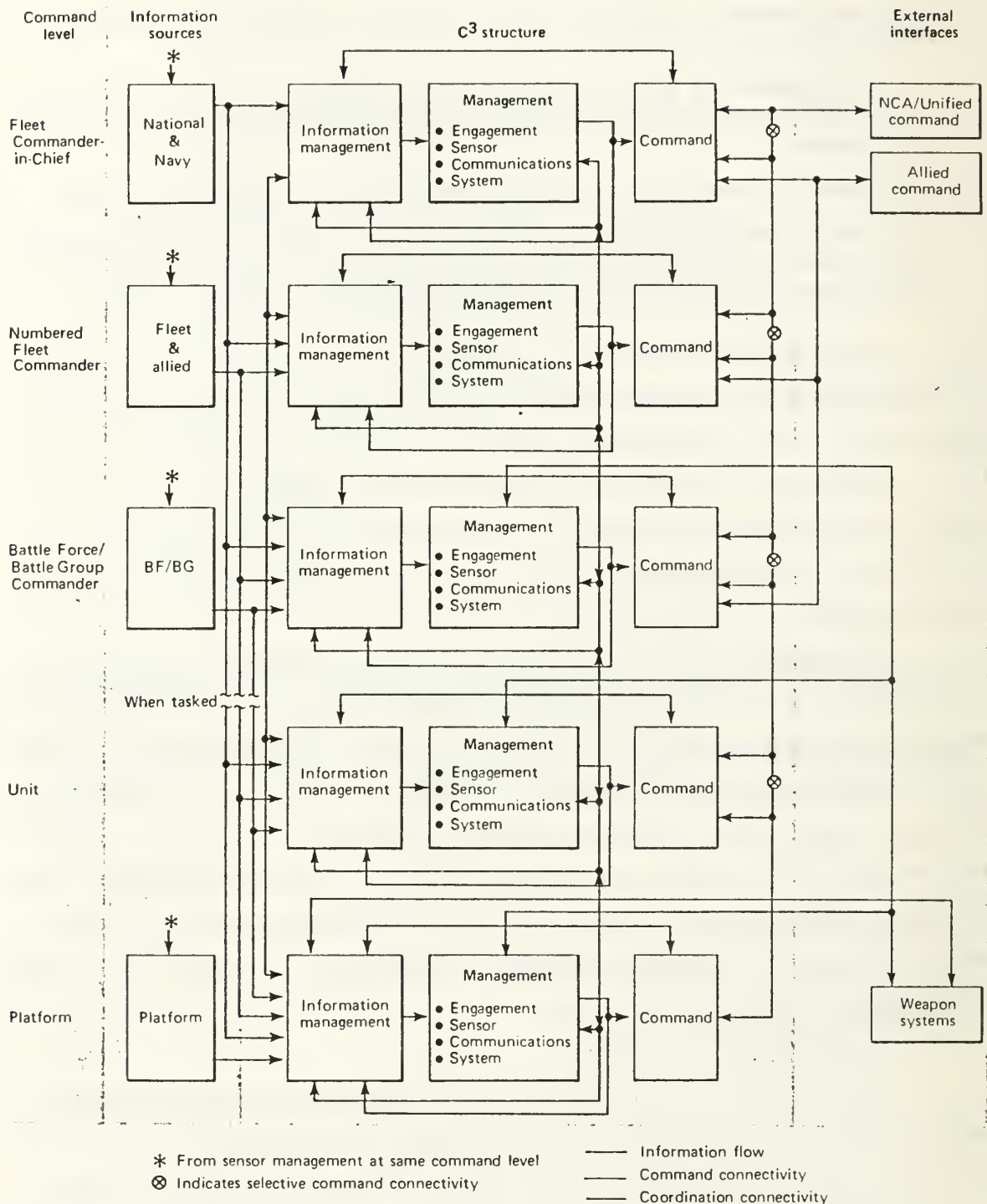


Figure 7. Conceptual C3 Naval Architecture: Structural depiction [Ref 3: p.15].

Halushynsky and Beam claim that the major characteristics of the Information Management function are the existence of the information base at each command, the processing at each level both ashore and afloat and the evaluation of tactical intelligence. The information is usually acquired in a data format for ADP because of volume and time constraints. Its processing has to do with the location determination of all platforms of interest and their subsequent classification.

Surveillance assets at each command level serve as information sources for air, surface, and subsurface contacts. The data is sent to the Information Management areas at own level and below. This goes down to the BG level for its area of interest on an equal basis and to the unit and platform levels whenever tasked. On the other hand, the exchange of processed information (the tactical picture) can take place both upward or downward, tailored to the interests and responsibilities of the intended recipient. This parallel connectivity between the Information Management areas allows the Fleet Commander-in-Chief to acquire the necessary data directly from any platform under his control. This connectivity ensures timely and accurate information which is essential in light of the highly sophisticated future threat.

Weapons systems can be controlled by the Engagement functional areas of the own level command or by the BG or unit command in case of a coordinated defense or attack. Targeting information is supplied from the Information Management areas of the respective command levels or by platform's own sensors. Finally unit commanders can control platforms appropriate to their assigned responsibilities. For example, the AAWC (Anti-air Warfare Coordinator) is responsible for the defense of the battle group against air attack. He can assign fighter interceptors from the BG assets as well as SAMs (Surface- to-Air Missiles) on various platforms to specific targets. The fire control requirements including midcourse guidance are the responsibility of each combat asset. Figure 8 summarizes these concepts covering part of the architecture down to the BF/BG command level.

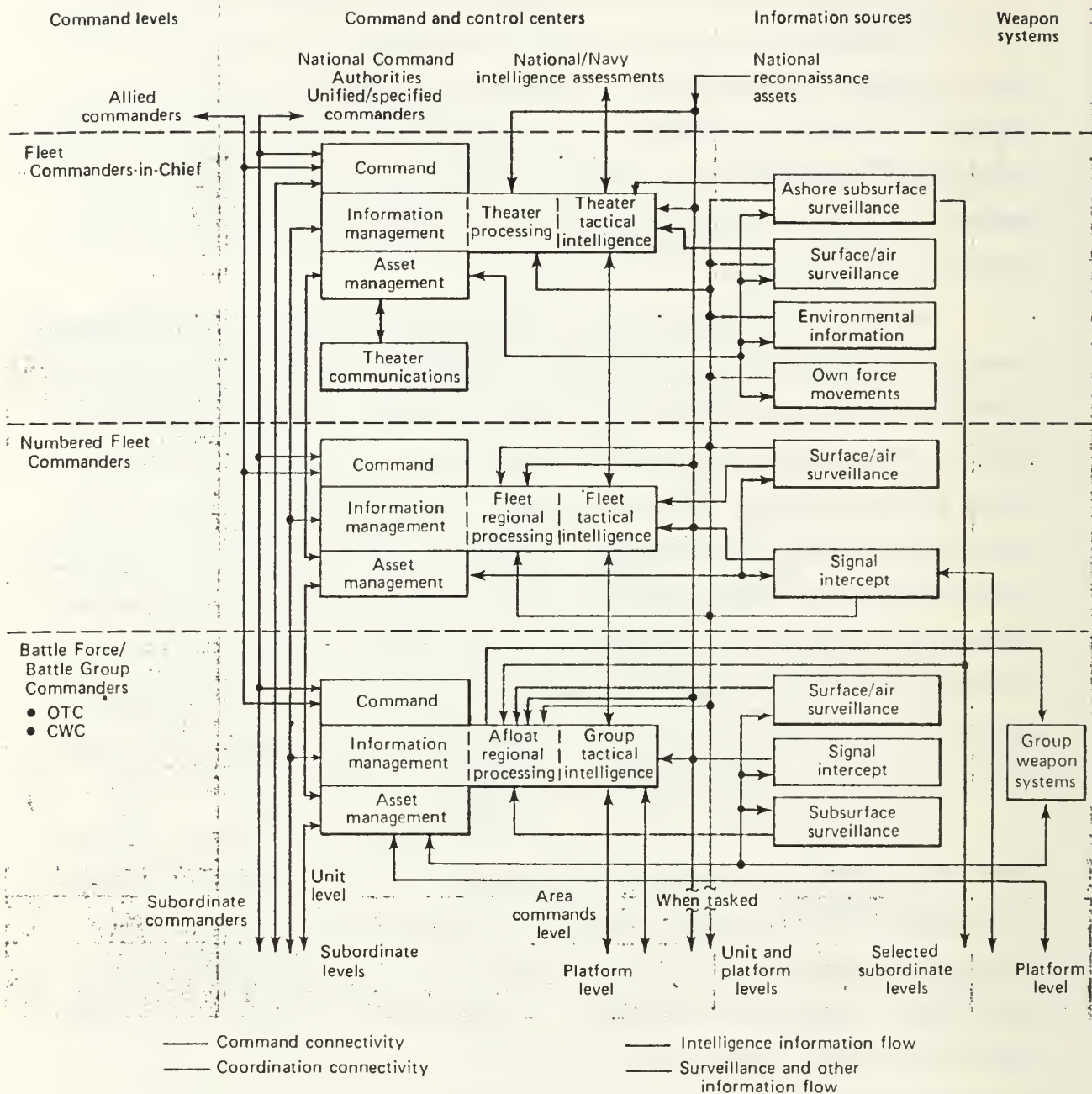


Figure 8. Conceptual C3 Naval Architecture: Partial structural depiction [Ref. 3: p.17].

F. OPERATIONAL/TECHNICAL CONSIDERATIONS

1. The requirements statement.

So far, a typical C3 naval architecture has been conceptually developed. Throughout this analysis the communications dimension appears to play a dominant and critical role. This role is even more profound given the complexity and the inherent limitations confronting a Naval Commander in a geographically restricted tactical area. He needs connectivity both with the environment and with his forces. He essentially needs an integrated tactical communication system to accomplish this task. This particular system must meet certain operational requirements. A thorough description of the features that the proposed system must possess includes:

- Autonomy.
- Survivability/Anti-jam capability.
- Security/LPI/LPE.
- Reliability/Maintainability/Availability.
- Compatibility/Interoperability.
- High Capacity and CNI functions.
- High Responsiveness.
- Connectivity.
- Ease of Installation and Operation.
- Extended Life Cycle and Affordability.

2. The requirements description.

The size, weight and power considerations of the system's terminals dictate the *autonomy* of the platform carrying them. These must suit the platform's space availability, without restricting its flexibility. Candidate platforms such as ships, fighter and reconnaissance/surveillance aircraft and command centers afloat have varying degrees of autonomy requirements. Furthermore direct data distribution from remote sensors to remote platforms enhances the degree of the system's autonomy since no central information collection would be necessary. Loss of a control center must not impair the system. Individual platforms of the system must be able to communicate as long as other termi-

nals exist. Complete availability and functionality is required to cope with current and future requirements. Overall the system must be flexible enough to adapt readily and quickly to any technological, organizational and threat changes.

Non-nodal connectivity is required to provide absolute system *survivability*. The wide platform dispersion and the information distribution independently of any network configuration will enhance the assets' survival. Continuously shrinking reaction times, especially in closed sea areas of operations where strike ranges are relatively short, have made the requirement for a survivable system essential. EMP (Electromagnetic Pulse) hardening is considered necessary for sealing the system's terminals. ARMs (Anti-radiation missiles), active enemy jamming and deception of system's transmissions can reduce the effectiveness of the system if this doesn't include advanced ECCM (Electronic Counter-Counter Measures) features for high *resistance to jamming*. Such enhancements are possible today with the use of state-of-the-art technology. Spread spectrum modulation and adaptive antenna nulling techniques form the basis for a solution which will guarantee transmission security (TRANSEC) even if the desired signal-to-jamming ratios are not favorable. In general, the system must be survivable enough to continually perform essential Command and Control functions irrespective of potential attacks.

Message security (COMSEC), is required through built-in encryption modules to provide low probability of exploitation (LPE). Low probability of intercept (LPI) must be also provided to counter enemy efforts for signal analysis through passive ESM (Electronic Support Measures) techniques. The system must be secure enough to allow naval functions to be undertaken without fear of compromise.

Enhanced *reliability* with minimal MBTF (Mean-Time-Between-Failures) should also be provided by the system. Extensive testing during procurement and built-in testing capability are needed for this purpose. Equipment commonality and a modular approach are needed for an *easy-to-maintain* solution.

Absolute *availability* is required for maintaining connectivity. Sufficient Mean-time-to-Repair (MTTR) is a must. The employment of such a system must provide terminals which will be technologically compatible with the existing equipment used to serve parts of the purpose for which this system is proposed. These are UHF/AM radios and hardware used for other tactical links. *Interoperability* among terminal platforms and among services is a must for providing maximum flexibility in future upgrades and ease of transition from the previous level of technology to the currently proposed one. Ultimately the goal here is to allow the naval forces and the naval commanders to interface actively with the forces and the commanders of other services.

Since *capacity* is driven by the number of nets handled by the system, it must be high enough to cope with the ever increasing needs for higher info-rates. An expanded wideband service can meet the communication requirements' growth and the accelerating number of links needed. This can be coupled with Multiple Access techniques offered by current technology, since the UHF band, a primary candidate for system's operation, is already congested. An integrative form of *Communication, Navigation and Identification* (CNI) capability must be addressed as well. Digital data link and precise navigation and identification functions must be incorporated into the system. The traditional TACAN (Tactical Air Navigation) and IFF (Identification of Friendly and Foe) functions should be performed by this system and in such a way that mutual interference and position ambiguity will be resolved.

Real-time information transfer is needed for operationally useful solutions. The system must be highly *responsive* to cope with this need.

Subscribers' *connectivity* should not be limited. No major space changes must be required for the *installation* of the system's terminals and no extensive *training* must be necessary for the use of the system. Having described the system's features, the next step is the initiation of the engineering effort which will essentially turn these features into hardware itself.

II. THE THEORY OF ANTI-JAM (AJ) SYSTEMS

The survivability of a modern integrated tactical communication system in an intense EW environment depends heavily upon its degree of protection against enemy intentional jamming, its level of signal undetectability and message coding efficiency. A current AJ system should possess such capabilities to meet these requirements in the most efficient way. The first two requirements can be currently provided by the employment of spread spectrum modulation techniques in the design of such a system. These techniques which are essentially 'internal' means of signal protection will be analyzed in detail in the following sections. Several existing 'external' techniques for signal protection, such as adaptive antenna steering, make use of the current technology by operating on the directivity parameter, in order to provide external signal protection. Although this further enhances the AJ capability of a communications system, it is not always desirable, as it happens here, with this proposed AJ system. Because this is essentially a broadcasting system, omnidirectional coverage is desired. Therefore, directional transmission is not considered. Thus, a brief discussion of adaptive null forming is included at the end of this chapter. The last requirement of message protection is normally met today by the employment of 'state-of-the-art' ciphering schemes, such as Reed-Solomon and Viterbi coding and decoding. Code theory is beyond the scope of this study and therefore it will not be analyzed here.

A. SPREAD SPECTRUM MODULATION TECHNIQUES

The means available to protect the transmitted signal are essentially radiated power, bandwidth and redundancy in the message. The most significant parameter in the protection of the message is the processing gain ratio, between the transmitted band and the signal band. Given a transmitter and a jammer, the loci of the points giving the same signal to noise ratio are circles whose radii depend upon the processing gain. This means that low interfer-

ence of the signal against jamming effects can be achieved by acting upon the signal processing gain.

From the operational point of view it means that the effects of the jammer are reduced or that the enemy is forced to radiate considerable power over a wide-band when the signal is radiated on a narrow band with 'frequency hopping' or continually when the signal is radiated pseudo-randomly in time.

Spread spectrum modulation is a method of increasing the bandwidth of a radio frequency (RF) carrier beyond the bandwidth usually required for it to be transmitted, and thus increasing the processing gain. The basic concept is simple. The actual information signal is introduced into the previously mentioned RF carrier, after being modulated in a digital form. If the intelligence is analogue as voice is, it is converted to digital form by 'vocoding'. This modulated signal is then mixed with the spreading pseudo-random code signal to create a wideband pseudo-noise signal for transmission. This final signal has a low power density and cannot be easily distinguished from random thermal noise.

This particular spreading of a modulated RF-carrier over the largest frequency range, possesses a number of considerable advantages: [Ref. 4: p.7-8].

- Selective addressing capability.
- Code division multiplexing for multiple access.
- Low density power spectra for signal hiding.
- Message screening from eavesdroppers.
- High resolution ranging.
- Interference rejection.
- Secrecy of communications.

The art of expanding the bandwidth of a signal, transmitting the expanded signal, and recovering the desired signal by despreading the received spread spectrum into the original information bandwidth is a process attempting to deliver error-free information in a noisy signal environment. The techniques utilized for this are the following:

- Direct modulation with pseudo-random sequence (PN).
- Frequency hopping (FH).
- Time hopping (TH).
- A combination of the previous three techniques (hybrid).
- Chirp modulation.

Figure 9 on page 23 shows a pictorial representation of the spread spectrum techniques in the time/frequency domain.

1. General Concept

The concept of the spread spectrum theory is expressed by C.E. Shannon in the form of channel capacity [Ref. 4: p.5-6]. The idea of increasing the security of a radio link using an artificially increased bandwidth is a direct result of his examinations regarding the behavior and the capacity of a telecommunication channel interference conditions.

$$C = W \times \log_2(1 + S/N)$$

Where,

C: Capacity in bits per second

W: Bandwidth in hertz

N: Noise power

S: Signal power

The above formula after mathematical simplifications can be rewritten in the form:

$$W = \frac{N \times C}{S}$$

The first formula shows the relationship between the ability of a channel to transfer error-free information, compared with the signal-to-noise ratio

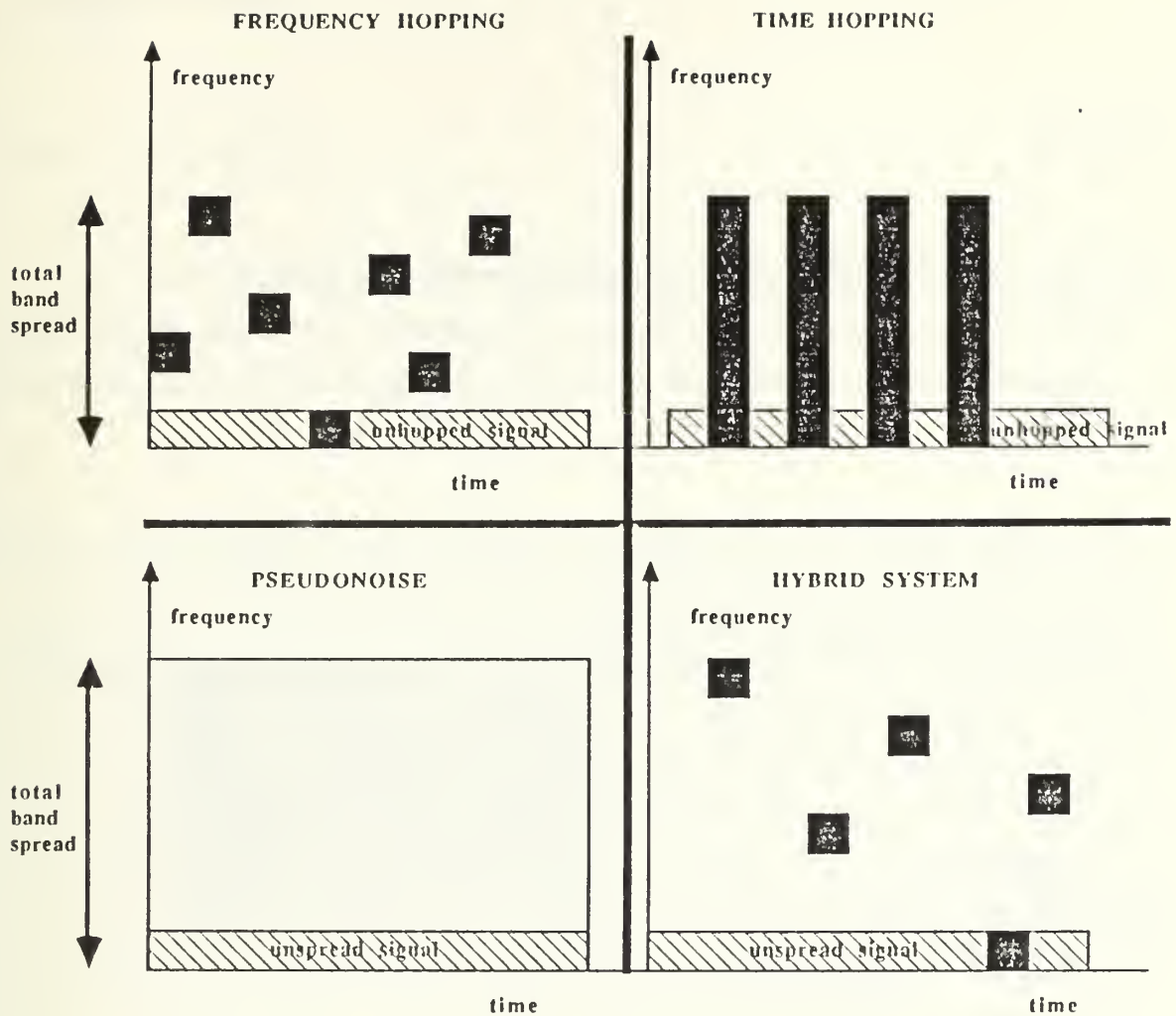


Figure 9. Spread Spectrum Techniques: Time/Frequency domain representation.

existing in the channel, and the bandwidth used to transmit the information. The second formula shows that for any given noise-to-signal ratio we can have a low information error rate by increasing the bandwidth used to transfer the information. The methods with which the information itself may be embedded on the spread spectrum signal will be examined later. Supposing that Transmitter Power is (S), Jammer Power is (J), Channel Bandwidth is (W) and Information bit-rate is (R).

It can be assumed that the power density N_0 of the interfering signal is:

$$N_0 = \frac{J}{W}$$

and the received energy per bit of the required signal E_b , is:

$$E_b = \frac{S}{R}$$

which describes that it is inversely proportional to the bit error rate. Rearranging the equation, it yields a new quotient J/S , as the ratio of interference power to signal power. This ratio represents the *jamming margin* of the system;

$$\frac{J}{S} = \frac{\left(\frac{W}{R} \right)}{\left(\frac{E_b}{N_0} \right)}$$

and it is the ratio by which a certain bit error-rate will be maintained.

With conventional systems, the bandwidth W of the transmission channel is equal to the bit-rate (SSB) or a multiple of this (AMDSB) so that the ratio (W/R) is constant. In these cases, the jamming margin is determined by the signal to noise ratio rather than (W/R) . With the spread spectrum techniques the ratio W/R is not constant. The bandwidth W can be artificially expanded with respect to the minimum bandwidth determined by the bit-rate R of the information. It is easily assumed that the jamming margin will increase linearly with the ratio W/R with a constant bit error-rate. The *processing gain* G_p can then be defined as the ratio W/R .

$$G_p = \frac{W}{R}$$

$$G_p(dB) = 10 \times \log\left(\frac{W}{R}\right)$$

The jamming margin of a complete telecommunication system can be calculated by taking into consideration the system losses V_{sys} , and the signal-to-interference ratio V_d at the demodulator. In this case the jamming margin is:

$$\frac{J}{S} = G_p - V_d - V_{sys}$$

A typical processing gain curve is shown in Figure 10. The increase of the jamming margin is possible, as the jamming threshold is dependent on the spreading code. For instance, the spreading of a narrow-band signal from 5 KHz to 50 MHz, increases the jamming threshold, to 36 dB, when the required S/N at the demodulator is +1 dB and $V_{sys} = +3$ dB. That means that the interference signal must be 36 dB stronger (approximately 4000 times) than the required signal, at the receiver input, before the demodulation of the required signal is affected. Thus, a satisfactory value for the processing gain is determined by high spreading code rates [Ref. 4: p.10].

a. The Direct Sequence Method

Direct sequence modulation is the modulation of a carrier by a digital code sequence whose frequency is much higher than the information signal bandwidth. This code sequence is a fast pseudo-noise (pseudo-random) one, which has a very wide spectrum with characteristics similar to the noise. Thus, the already modulated and amplified RF carrier is increased in bandwidth by using BPSK modulation with a certain, suitably binary code sequence of pseudo-noise type (PN). It can be achieved by using a balanced modulator for the PSK, such as a four-diode mixer. This mixer is a 0/180 phase shifter. The whole process and the respective spectra of the individual signals are

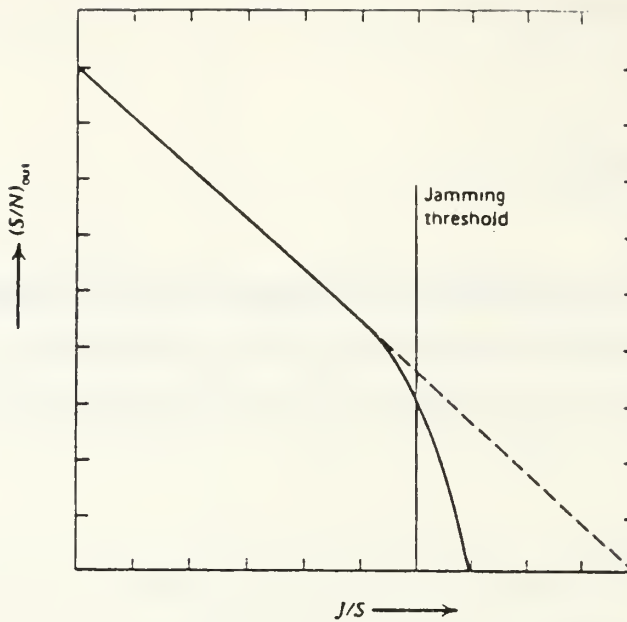


Figure 10. Typical Processing Gain Curve: jamming threshold [Ref. 4: p.12].

shown in Figure 11. The BPSK balanced modulator is analyzed in [Ref. 4: p.109-125]. The actual information can be embedded either on the RF carrier (e.g. FM modulated RF carrier), or on the PN code sequence by modulo 2 addition. See Figure 12.

The PN sequence used for spreading the signal has a frequency spectrum whose envelope has the function $[\sin w/w]^2$. The zero positions correspond to multiples of the clock frequency. The spectrum of the phase shift carrier also is of the same shape $[\sin w/w]^2$, which is located symmetrically around the original carrier frequency. Since its higher power density is concentrated into the main lobe, it is sufficient, in order to process this signal, to accept, as the minimum required bandwidth, the frequency-band between the first two null-to-null points around the center frequency. This bandwidth is determined by the code-rate and it is identical with twice the clock frequency of the code sequence. See Figure 12.

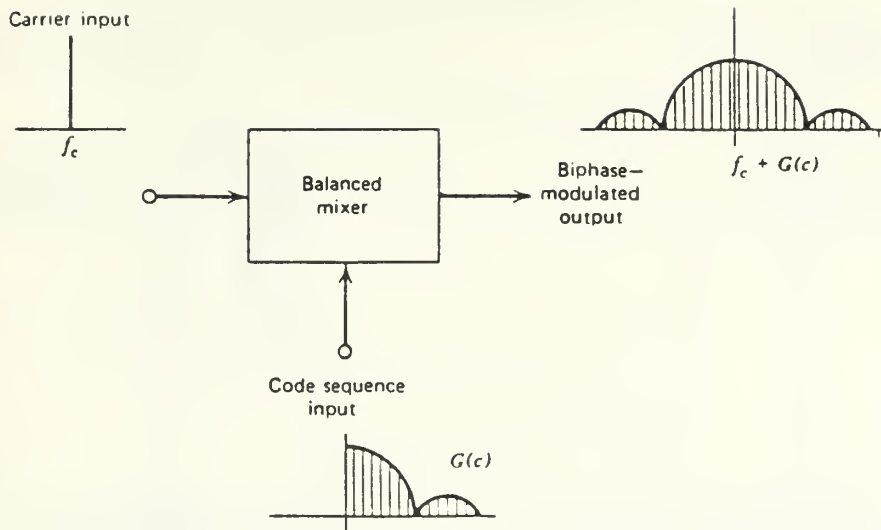


Figure 11. Direct Sequence (DS) Modulator: frequency spectrum representation [Ref. 4: p.17].

The main characteristics of this particular power density spectrum have been studied and verified in the laboratory. These are listed below [Ref. 5: p.6].

- The Bandwidth of the main lobe is identical with twice the code rate.

$$BW_{ss} = 2 \times R_c$$

- The spacing of the spectral lines is determined by the code rate and the code length (maximal) [Ref. 4: p.28].

$$\text{Spacing: } \left[\frac{R_c}{2^n - 1} \right]$$

- A suppressed carrier signal, produced by employing biphase-balanced modulation, is difficult to detect because it is well below the noise level produced by the code modulation.
- More power is available for sending useful information because the transmitter power is used to send only the code produced signal.

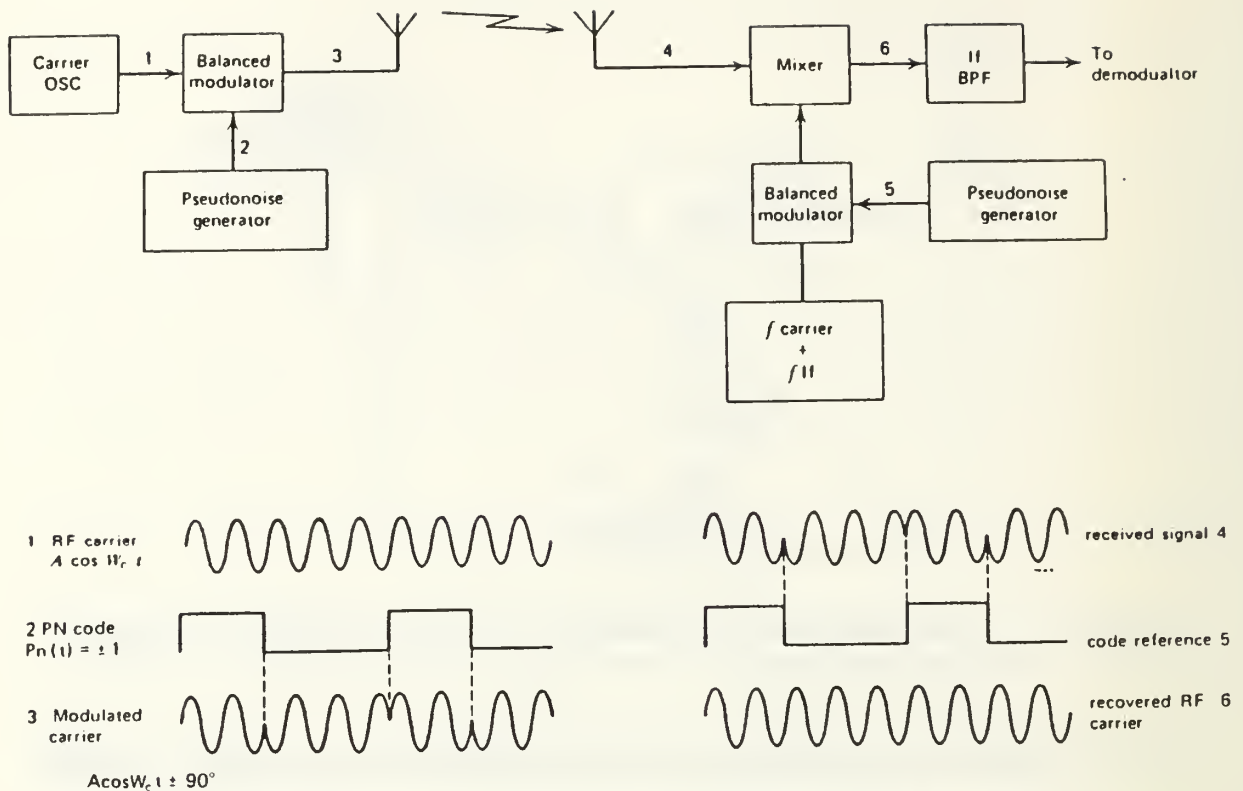


Figure 12. Direct Sequence (DS) Tx/Rx: time domain representation [Ref. 4: p.18].

- The signal has a constant envelope level so that transmitted power efficiency is maximized for the bandwidth used.
- The power of the main lobe is determined by the code rate, and it contains almost 90% of the total signal power. High side lobe power density is not useful (see Figure 13).

$$\text{Power density: } \frac{1}{2 \times R_c}$$

- In any case the direct sequence system has a main lobe bandwidth that is the function of the waveshape and the code-rate used (see Figure 13). The power spectrum of a purely random binary code sequence with a bit duration T_0 is derived in [Ref. 6: app. E].

Generally, with sufficiently large spread bandwidths and suitable code sequences which have a virtually random character, the RF power of the

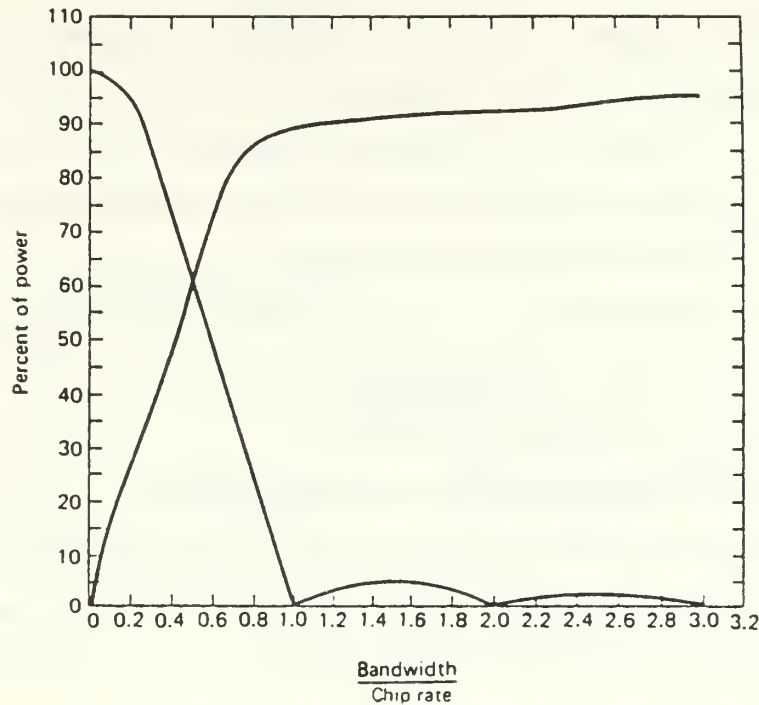


Figure 13. Direct Sequence (DS) Spread Spectrum: power distribution [Ref. 4: p.23].

original carrier will be distributed virtually constantly over a wide frequency range, and the power density (W/Hz) will have a very low value. When received on a narrow band receiver (not synchronized and authorized), this signal will be heard as noise only, which will hardly change when turning over a wider frequency range, and which can be so weak that it disappears into the interference and noise level [Ref. 5: p.7]. Such transmissions are very secure and difficult to discover, and they will hardly ever cause interference to other telecommunication links using the same frequency range, due to their low spectral power density. The prerequisite is, of course, a sufficiently high carrier suppression of the mixer, and good linearity of the amplifier so that the carrier is not regenerated due to intermodulation.

The military application of this method is obviously based on these unique properties. Furthermore, by exchanging bandwidth for signal-to-noise ratio according to Shannon's information theory, the required communication

capacity can still be achieved with a consequent saving in power [Ref. 6: p.76], (keeping S/N low-increasing W/R). At the receiver, the signal is correlated with an exact replica of the PN code sequence used at the transmitter. The phase reversals of the original PN code are thus removed in this de-spreading process, while the phase reversals due to the binary information alone are then extracted by a PRK demodulator (see Figure 14). From above, it is apparent that very many characteristics of the transmitted spread signal must be already known at the receive end (authorized receiver):

- The code sequence used for spreading.
- The exact clock frequency of the spreading code.
- The exact starting time of the code (phase position) (synchronization).
- The exact carrier frequency (carrier and clock frequency can be derived from a standard synthesizer at both ends).

The correlator mentioned above is again a biphasic balanced modulator which serves as first mixer by using an already spread RF-carrier that has been shifted to the value of the IF.

The encoding signal for the data information is usually a maximal length pseudo-random code sequence derived from an n -stage shift register using feedback. The length of the codeword is $2^n - 1$ bits after which it repeats itself. The shift register is clocked typically by a frequency source at a high bit-rate (e.g.: 1Mbit/s). The output of the shift register is then combined with the data stream using modulo 2-addition by means of logic gates (see Figure 12). In practice, there are a number of such code "patterns" which may be generated for multichannel operation and the most suitable codes are those which yield the minimum amount of cross-talk due to cross-correlation at the receiver.

(1) *Processing Gain of DS.* The processing gain of a direct sequence system is a function of the RF bandwidth of the transmitted signal compared with the bit rate of the information (R_{info}).

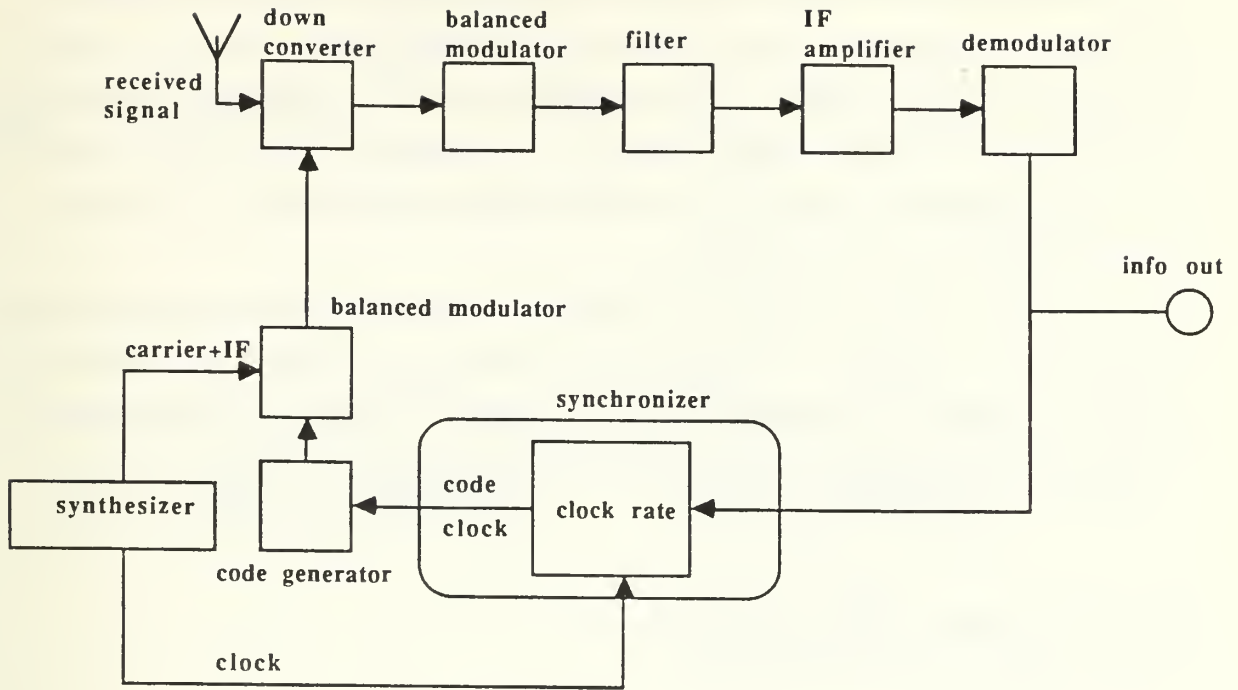


Figure 14. Spread Spectrum Direct Sequence Rx: block diagram.

$$G_p = \frac{BW_{RF}}{R_{info}}$$

The assumption is that the RF bandwidth is the same as that of the main lobe which is twice the bandwidth of the spreading code clock rate. Given that normally only 90 percent of the total power is contained in the main lobe, for a system having a 10 Mcps (Mega chips per second) code rate and a 1 Kbps information rate the processing gain would be:

$$G_{pdB} = 10 \times \log \left[.9 \times \frac{10^7}{1 \times 10^3} \right] = 39.54dB$$

It is assumed that the processing gain is only limited by current technical capabilities. Available integrated circuits allow limited code generation at rates up to 200 or even 300 Mcps. As the code rate increase, operating errors have more impact on the overall performance. Also at higher rates it is not desirable to have too much gain. The practical upper limit tends to be between 50-100 Mcps.

On the other hand, there is a lower limit to the data rate, due to propagation instability and phase noise at the local oscillator. A typical limit of 10 bps is used. Now, considering the two marginal limits we can see that for practical systems the processing gain G_p is limited to about: [Ref. 4: p.26].

$$G_{pdB} = 10 \times \log[.9 \times \frac{10^8}{10}] = 69dB$$

(2) *Applications of DS.* In conclusion, the main application of the spread-spectrum direct sequence method is in the field of satellite communications. As satellite transmission power is very costly to provide, it must be used effectively. It is also used in hybrid techniques in combination with 'Time Hopping' or 'Frequency Hopping'. Despite some of their vulnerabilities, direct sequence systems are extensively used because of their simplicity and low cost, in providing high G_p .

b. The Frequency Hopping (FH) Method

The second method of spreading the frequency bandwidth of a transmitted signal is to allow the carrier frequency to jump from one frequency to another by selecting one of a large number of available channels. This sequential selection of the frequencies is made according to a PN-code sequence, and will therefore appear to be random in a non-authorized receiver (See Figures 15, 16).

The authorized receiver controls its local oscillator with the same PN-code sequence and will therefore follow the frequency hopping pattern of the transient frequency with the correct clock and phase. This particular

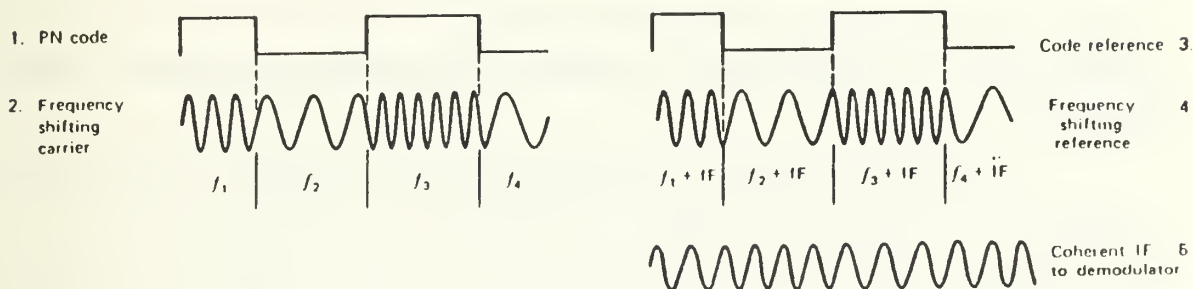
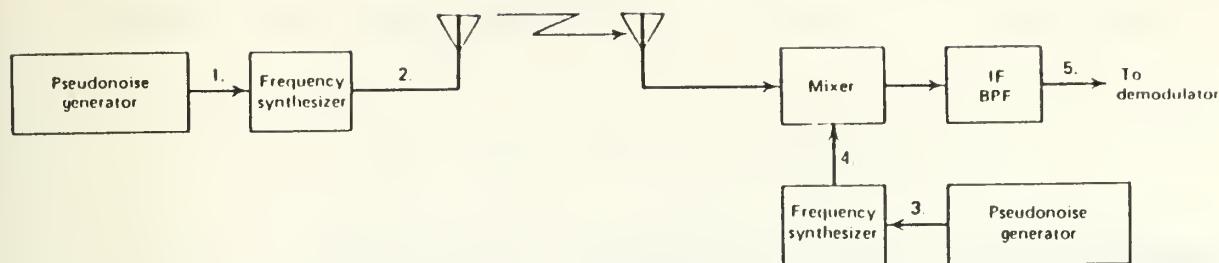


Figure 15. Frequency Hopping (FH) System: time domain representation [Ref. 4: p.30].

method can be easily regarded as the present state-of-the-art, and it is feasible since fast frequency synthesizers and the required programmable dividers, PLL-circuits, and VCOs are available. The theoretical processing gain that this method gives is :

$$G_p(FH) = \frac{BW_{RF}}{R_{info}} = N$$

Where N is the number of the available channels.

For example, for N = 1000 channels among which a hopper jumps back and forth, the processing gain will be : $10 \times \log 10^3 = 30dB$ [Ref. 4: p.31]. The PN code sequence which determines the sequential "hops" within the

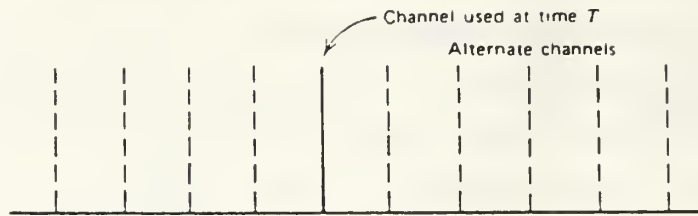


Figure 16. Frequency Hopping (FH) System: frequency spectrum representation [Ref. 4: p.31].

range available, is created as in the case of DS method, by using the digital data stream.

Frequency hopping is an ECCM technique originally conceived as a method of combatting the effects of communications jammers. It also provides LPI against enemy ESM activities such as direction finding and position location. The main task of this method is to evade the jammer- not resist it. To achieve this, it is essential that the hopping pattern be unpredictable to the hostile forces. This is met by using the PN code. The performance of frequency hopping against various types of jammers will be discussed in the following sections.

(1) *Frequency Hopping Rate Determination.* The type of information being sent and its rate, the amount of redundancy used, if any, and the distance to the nearest potential interferer, are the main factors that determine the minimum frequency switching rate in a frequency hopping system [Ref 4: p.34].

As previously mentioned, a large number of channels, usable on demand, gives a high G_p for the particular frequency hopper. This can be seen easily in the following description, and digital modulation is again to be assumed for simplicity.

One channel is to be interfered with during these considerations, where the interference power should be greater than that of the required power. In a conventional narrowband system, the error rate in this

channel would then be 1 or 100%. In the case of a FH-system, the error rate amounts to the following, if one data bit is transmitted per frequency jump:

$$\text{No of interfered channels/No of available channels} = J/N$$

A narrow-band interfering station with $J=1$ would result in an error rate of $p = 1 \times 10^3$ in the case of a 1000-channel FH system with $N = 1000$. This is already a good value for voice communication; however, it is not sufficient for data transmission. For this reason, one has started to use more than one frequency jump per data bit (e.g. 3 per data bit), and to make a majority decision at the receive and (e.g. 2 from 3). Then the possibility of errors can be considerably reduced. If "c" represents the number of channels per bit, "r" the required number of interfered channels in order to interfere with one bit, and "p" the possibility of error for a normal FH system, the new possibility of error P is as follows:

$$P = \sum_{x=r}^c \binom{c}{x} \times p^x \times q^{(c-x)}$$

In the previous example with $J=1$, $N=1000$, $p = 1 \times 10^3$, the probability of error can be reduced to the following with $c = 3$ channels per bit and a 2-out-of-3 decision criterion:

$$P = \binom{3}{2} \times p^2 \times (1 - p)^{3-2} = p^2 \times (3 + p) = 3 \times 10^{-6}$$

The error rate versus the number of channels jammed (J/N) for various chip decision criteria in multichip transmission is shown in Figure 17 [Ref. 4: p.35].

The general characteristics of the frequency hopping method are then the following:

- The frequency hopping rate is a function of the information rate.
- The information error rate cannot be reduced below J/N unless some form of redundant transmission is used.
- The number of frequency channels required is determined by the desired interference rejection capability (jamming margin) and chip error tolerance.
- The correlated (dehopped) signal seen in a non-coherent frequency hopping receiver follows the shape of the function below:

$$\left(\frac{\sin x}{x} \right)^2$$

- It has a main lobe bandwidth equal to twice the chip rate.
- Chip rate is bounded on the low side by multipath and repeater interference considerations.

The best rule for variable geometry situations is to hop as fast as possible. This will be discussed in more detail in the following sections. Again the synchronization between transmitter and receiver, will enable the receiver to follow the PN-sequence and hopping pattern in accordance with the transmitter, and as such is essential.

c. The Time Hopping Method

This method, used to counter impulsive jamming, is normally used in combination with other techniques as it is less efficient in terms of exploitation and deception. It consists of varying, in accordance with a PN code sequence, the time of signal transmission. It also widens the spectrum of a transmit signal and when it is used together with DS or FH systems can offer a considerable improvement to their characteristics. It is mainly a kind of pulse modulation; that is the code sequence is used to key the transmitter on and off. These on and off times are pseudo-random, like the code, with an av-

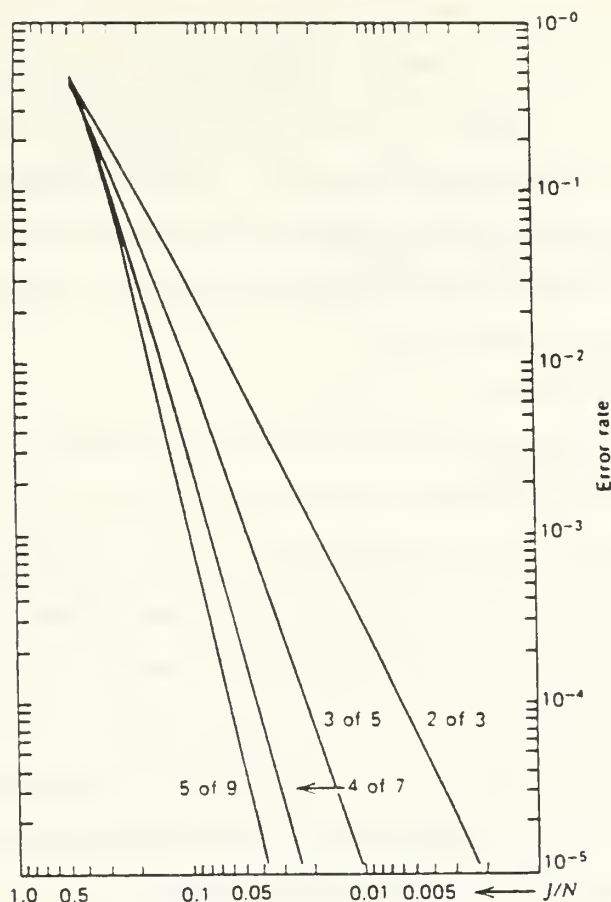


Figure 17. Frequency Hopping (FH) System: error rate versus fraction of channels jammed (J/N) [Ref. 4: p.37].

erage transmit duty cycle of about 50%. It is used for reducing interference between systems in TDM. Stringent timing requirements must be placed on the overall system to ensure minimum overlap between transmitters. Also as in any other coded communications system, the codes must be considered carefully from the standpoint of their cross-correlation properties.

This type of modulation offers very little in the way of interference rejection, as a continuous carrier at the signal center frequency can block communications effectively. The main advantage offered, is in the reduced duty cycle; that is, an interfering transmitter, to be really effective, it would be forced to transmit continuously (assuming the PN coding used is unknown to

the interferer). The power required of the reduced duty-cycle time hopper would be less than of the interfering transmitter, by a factor equal to the signal duty cycle. As was previously mentioned, when used in a hybrid form with FH it offers improved AJ performance against single frequency interferers. It is also useful for multiple access and other special uses, mainly because of its simplicity in generating the transmitted signal [Ref. 4: p.44]. A block diagram of TH Tx/Rx is shown in Figure 18.

d. Hybrid Forms

The combination of FH, TH and PN modulation result in some hybrid techniques, which are best used to ensure the required degree of protection against jamming and interception. They allow the adaption of the system characteristics to the requirements of the operational scenario, with simpler means. Furthermore, other signal processing techniques, such as forward error correcting codes, contribute to the protection of the message.

A hybrid DS/FH-system, employs the direct sequence and the frequency hopping method simultaneously. The total processing gain is equal to the product of the gains of the individual methods:

$$G_{p\text{total}} = G_p(DS) \times G_p(FH) = \left(\frac{W}{R} \right) \times N$$

This combination is very important, when a certain processing gain cannot be obtained using a single method on its own, or only with difficulty.

If, for example, a gain of 47 dB is required at a data rate $R = 10\text{Kbps}$ for interference suppression, it will be necessary to spread the signal over a bandwidth of $W = 500\text{ MHz}$ in the case of a DS-system, which will require a PN-clock rate of 250 MBit/s . In the case of a FH-system, it will be necessary to hop between 50,000 channels. Both possibilities are at the limits of what is technically possible. If, now, we take a combination of these two methods, it is possible using a DS-code of 10 Mbit/s and a 50-channel

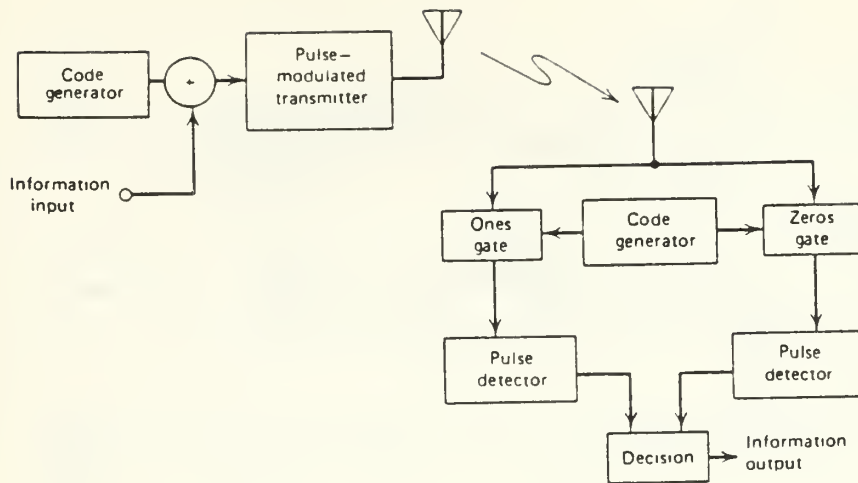


Figure 18. Time Hopping (TH) Tx/Rx: block diagram [Ref. 4: p.43].

FH-system to obtain the previous processing gain of 47 dB, with normal means [Ref. 4: p.5].

Typically, in spread spectrum communication systems with multiple access, the DS or FH-method is often used together with the time hopping (TH) method. In this case, the individual stations transmit only at several time intervals, determined by a common PN code. This allows us to achieve the condition that only one station is transmitting at any particular time. Even though spread spectrum signals can be operated quite well in the same frequency band, it is possible for local transmitters and receivers to be affected by unwanted desensitization. The block diagrams of several hybrid forms are shown in Figures 19, 20 [Ref. 4: p.47-54]. Some of their main characteristics are given in the next sections.

(1) FH/DS Modulation

- Hybrid FH/DS transmissions are straightforward super-impositions of direct sequence modulation on a frequency hopping carrier (see Figure 19).
- The same code sequence generator can supply code-data both to the frequency synthesizer to program its hopping pattern and to the balanced modulator (see Figure 19).
- The DS code rate is normally much faster than the rate of FH.

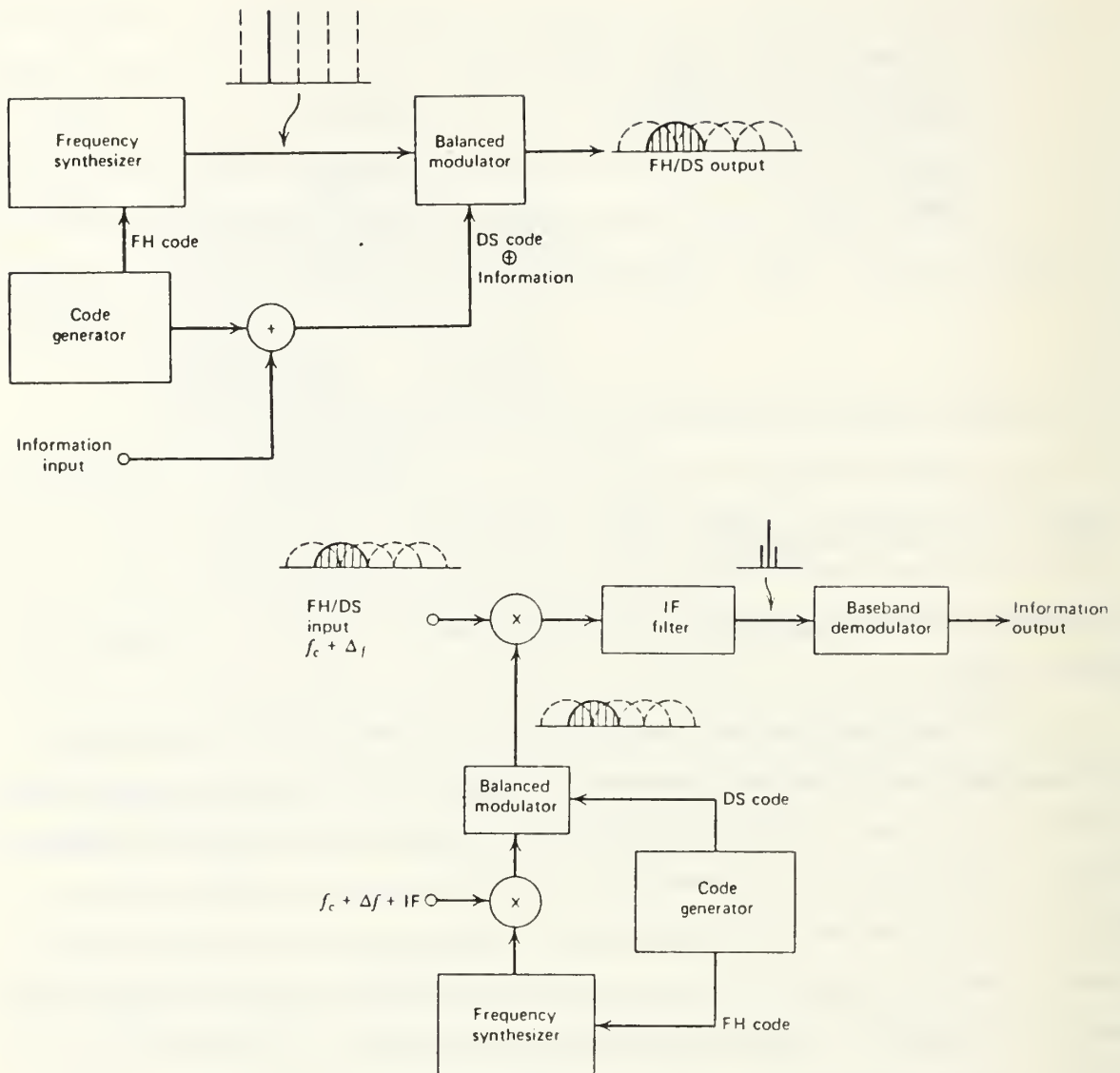


Figure 19. Frequency Hopping Direct Sequence (FH/DS) Tx-Rx: block diagram [Ref. 4: p.49-50].

- The jamming margin can be significantly less than the simple FH or DS system. Further explanation will follow.
- The composite processing gain is higher than that of the individual methods.
- This method extends spectrum spreading capability.
- Provides multiple access and discrete address.

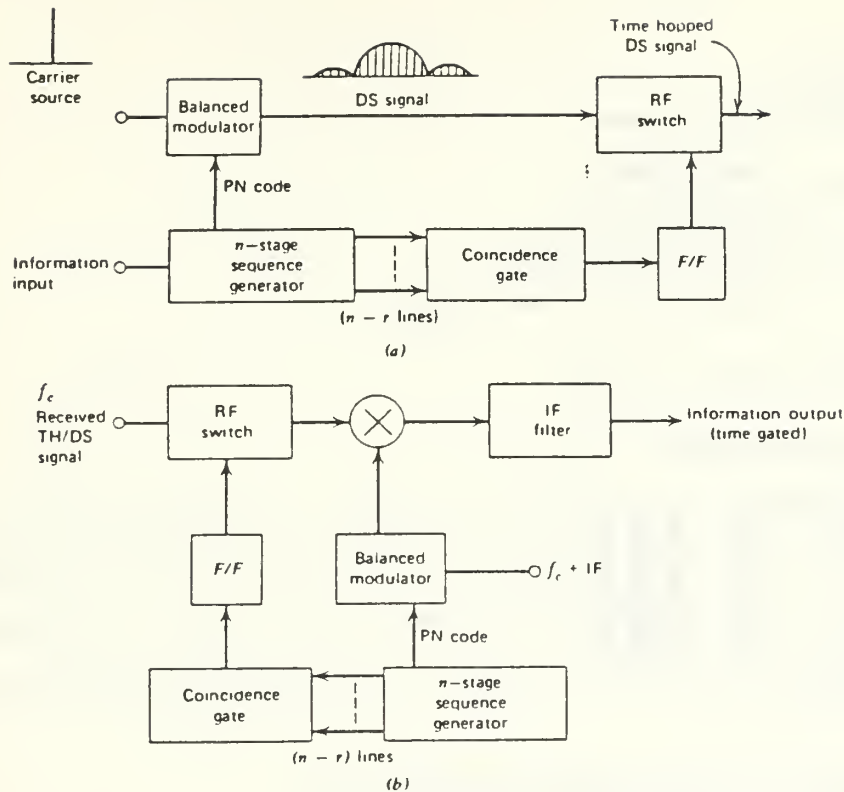


Figure 20. Time Hopping (TH/DS) Tx-Rx: block diagram [Ref. 4: p.54].

(2) TH/FH Modulation

- Hybrid TH/FH employs transmission and reception on several frequencies, at different times.
- Provides random access and discrete address.
- Improves anti-jamming capacity of spread spectrum systems.
- Reduces the 'near-far' effect.

(3) TH/DS Modulation

- Hybrid TH/DS employs time-hopped DS transmission in a pseudo-random manner (see Figure 20).
- Time hopping 'on-off' decisions can be easily derived from the same code sequence generator, deriving the spectrum spreading code.
- This pseudo-random time division permits more channel users, improves operation over simple code division multiplex and allows more accesses.

- Every DS signal 'on' period would occur at an average rate of:

$$\left(\frac{R_c \times 2^r}{2^{n+1} - 2} \right)$$

Where,

R_c: code clock rate

n: shift register length

r: register stages not sensed

(See Figure 20), [Ref. 4: p.54].

B. MULTIPLE ACCESS

A busy tactical communication link usually requires relays that are capable of conveying heavy traffic originating at many ground stations to a large number of receiving destinations. These requirements led to the development of multiple access techniques. The main methods of multiple-access are: (see Figure 21)

- Frequency Division Multiple Access (FDMA).
- Time Division Multiple Access (TDMA).
- Code Division Multiple Access (CDMA) or Spread Spectrum Multiple Access (SSMA).

The first two forms are exact analogies of the multiplexing schemes used in analogue and digital multichannel terrestrial systems. CDMA has found only restricted use in military systems.

Having achieved access to the relay with one of these techniques, then the second basic characteristic is the type of modulation used in the specific accessing channel. After this, a way of handling heavy traffic within the channel must be defined. The traditional method is the multiplexing of signals, which is a process in which a number of different messages are combined to be transmitted over a common communication channel. Figures 22, 23 show the categorization of these techniques and the multiple access signal flow respectively.

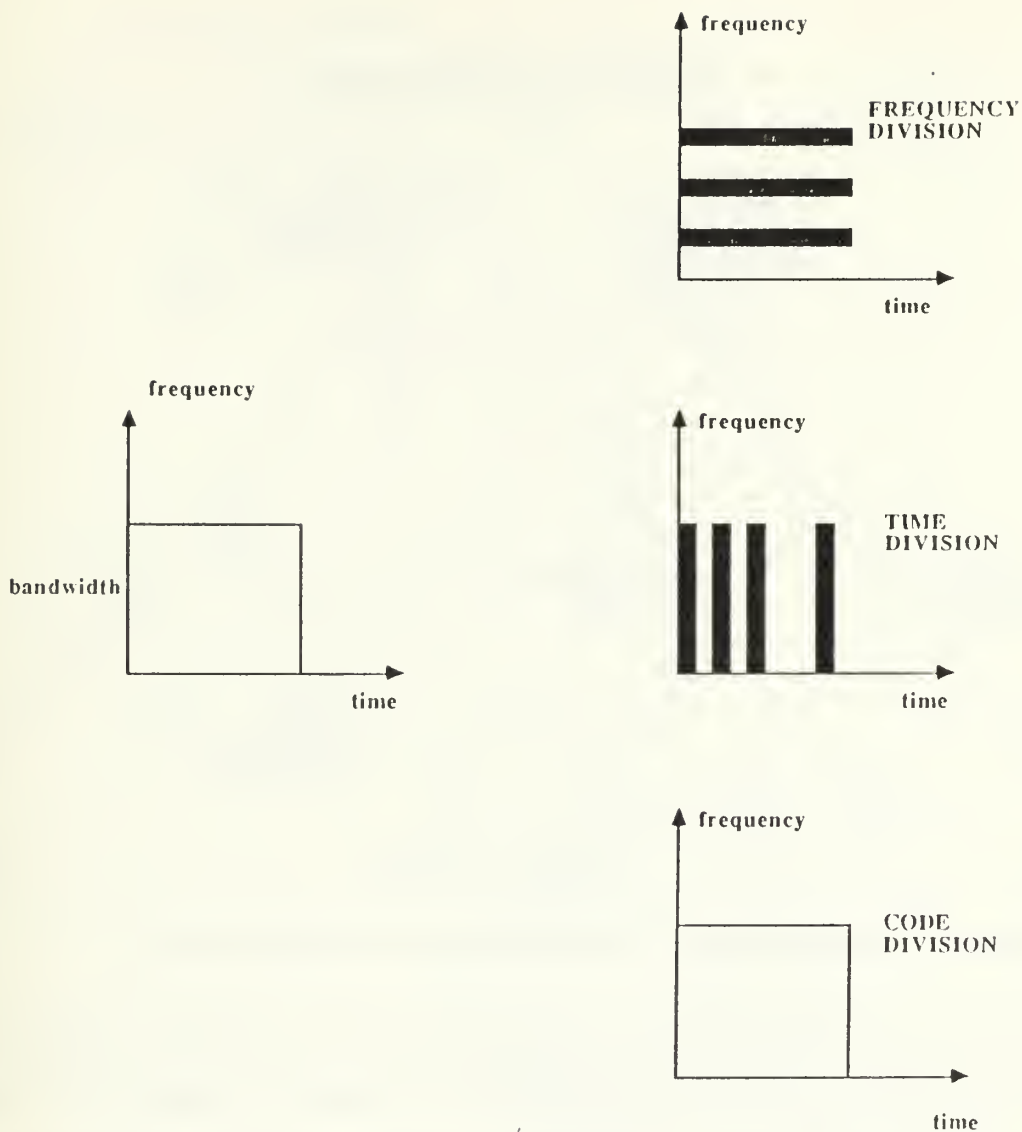


Figure 21. Multiple Access: Division of time-frequency space.

a. Frequency Division Multiple Access (FDMA)

In FDMA the earth stations are assigned frequency bands for the transmission of their signals. The assigned bands are separated by guard bands to reduce interference. The relay power is divided among the carriers that are accessing it all the time. The most common types of FDMA schemes are FDM/FM/FDMA and SCPC/FDMA (single channel per carrier). The first

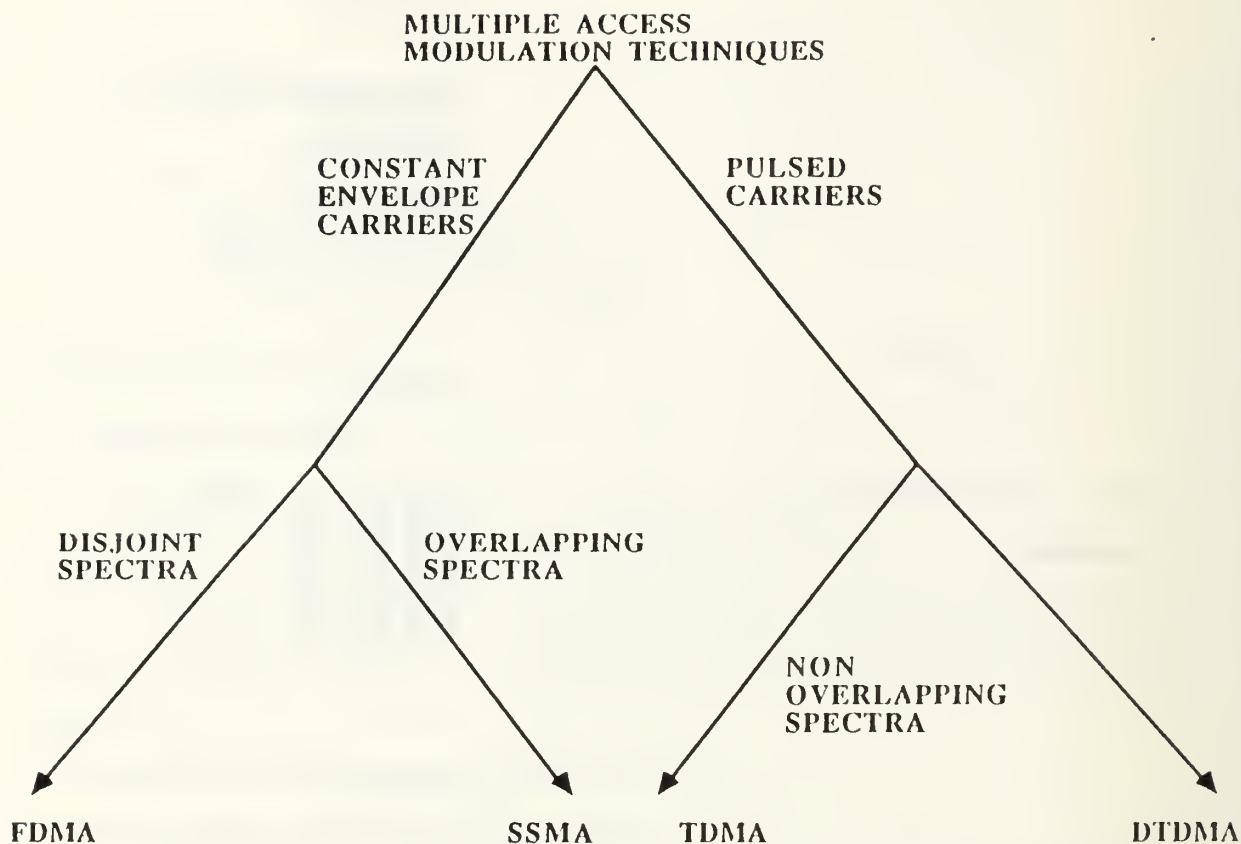


Figure 22. Multiple Access: categorization of several techniques.

technique was introduced in the analogue modulation and the second in the digital modulation. As the name implies, each carrier is utilized for a single analogue or digital channel, in terms of accessing the relay. In terms of modulating the assigned carrier within the channel, FM and QPSK are used respectively. For the multiplexing of several messages in each particular channel, FDM is also used. In FDM, the available spectrum of the channel is divided so that the signal at the transmitter is allocated to a different part of the spectrum, and the demultiplexing at the receiver is achieved by frequency selective filters. The disadvantages of this method, such as cross-modulation and inter-modulation products, affect its use.

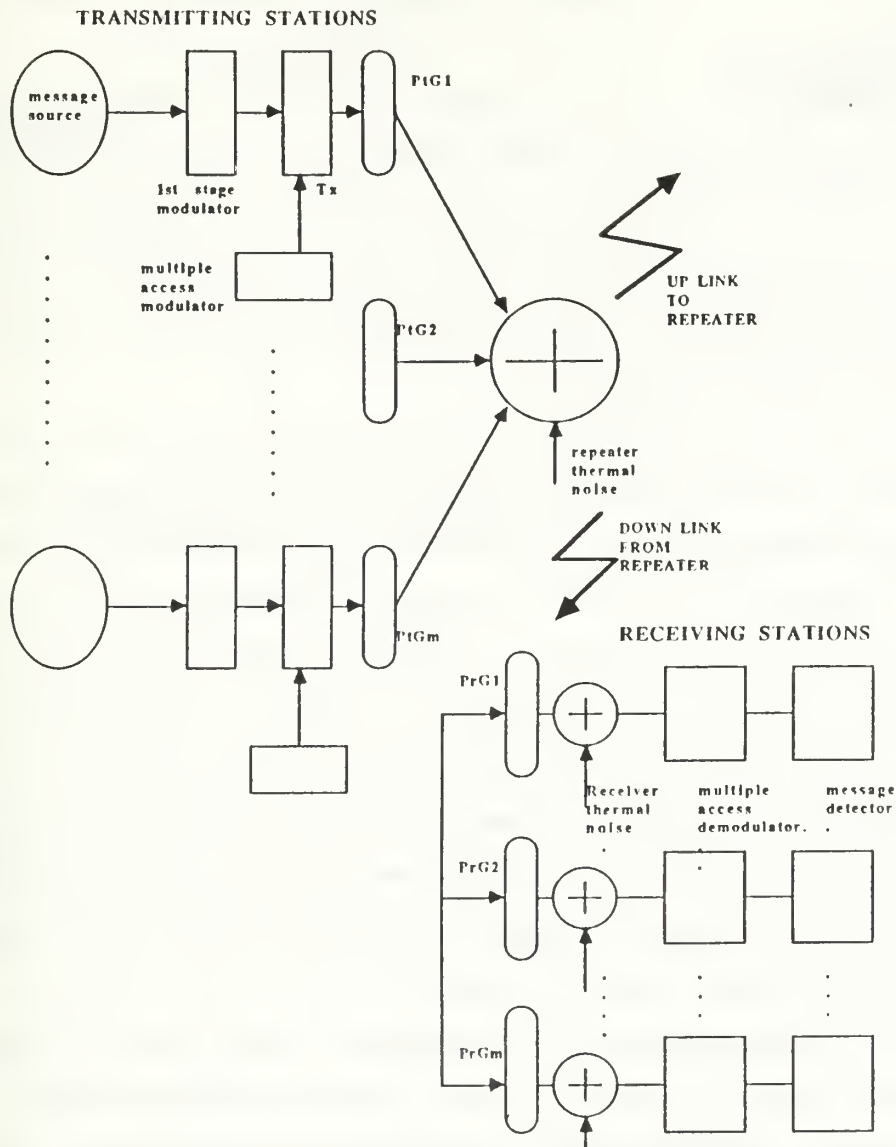


Figure 23. Multiple Access: signal flow.

b. Time Division Multiple Access (TDMA)

In TDMA, the earth stations are assigned time slots which are separated by guard time periods for the transmission of their signals. During its assigned time slot the earth station (transmitter) has the entire transponder (relay) bandwidth and power available. This type of access is a time gating

function which correctly locates each transmission burst relative to the rest of the network.

It is now clear that the signals pass through the relay with no interaction. An access channel in TDMA designates a particular sequence of time slots. These time slots occur periodically at a definite repetition frequency, known as the 'frame repetition rate' of the system. All other channels in this arrangement would have the same frame rate but different times of occurrence, and possibly slot durations.

It is also essential for this system to use a signal as a time reference or relay clock, for maintaining synchronization. So the signals can be kept separated in time and in a specified sequence by maintaining specified time intervals relative to the relay clock. This can be done to within $0.1 \mu\text{secs}$. The procedure of time is illustrated in Figure 24. Briefly, each station that wishes to transmit a signal to another station, makes use of the timing marks that are transmitted by the relay, at intervals of T seconds. Then to set its local clock, a station transmits a series of pulses at intervals nominally T seconds, in length. It then compares their repetition rate, when they are received after retransmission, it then adjusts its local clock PRF to the PRF of the received signal, such that the difference is eliminated.

The repeater bandwidth must be large enough to keep the signal in one time slot from affecting signals in other time slots. The user stations can transmit message information in any manner they desire, provided that the message waveforms are suitable for passing through the repeater and do not overlap their assigned time slots.

In determining the time frame length, it is assumed that there are two modes of operation. The first mode allocates each station one time slot per frame. For real time voice transmission, the frame rate could be equal to the convenient sampling rate (8 kHz), is $T = 125 \text{ ms}$. The second mode is a store-and forward operation in which many message samples are stored and transmitted together in bursts.

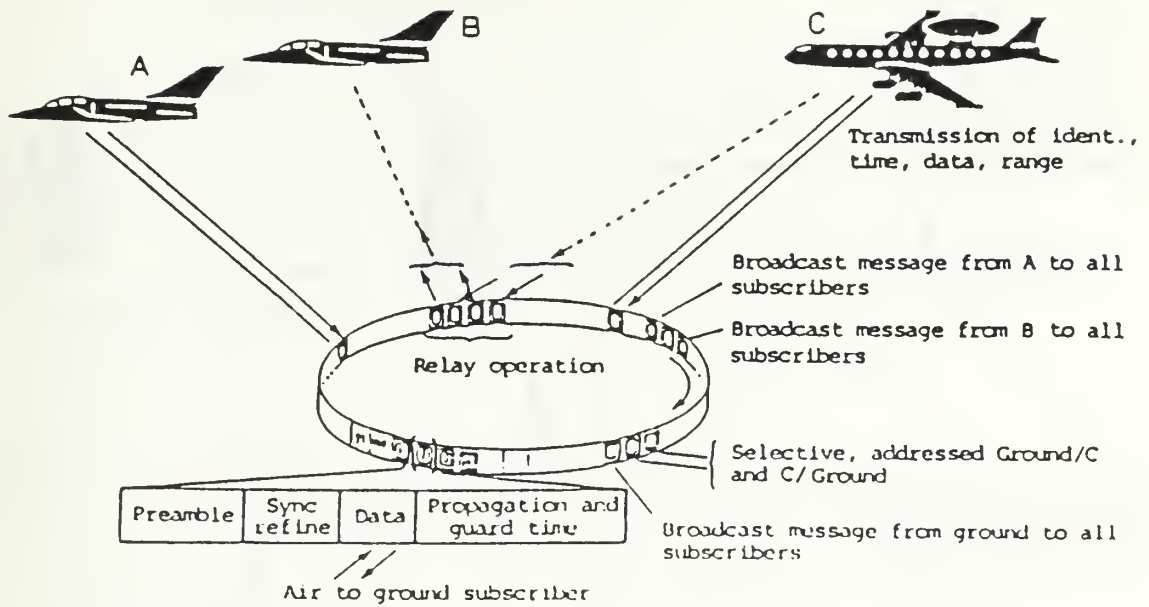


Figure 24. TDMA: data transmission.

(1) *TDMA frame and Burst format.* The typical TDMA and burst format is shown in Figure 25. For simplicity a four-access network is shown, but up to about 60 could be accommodated in an actual system. Typically 5 to 15 access per frame are expected in operating networks. A typical burst is expanded in the lower portion of the same figure. The guard time is usually 100 to 200 msecs, providing an assurance against successive burst overlapping. The burst lengths also are normally variable, since variable gross traffic loads are transmitted by each station. Reconfiguring of the traffic load is easily accomplished by reallocating the burst lengths assigned to each station, by some means of digital control. Then a carrier-recovery/ bit-timing-recovery (CR/BTR) sequence follows. This is used by the PSK demodulator in each receiver to recover local carrier or digital clock for coherent demodulation. For example, the CR/BTR sequence is typically 60 bits in a 60 Mbs system.

After CR/BTR, a unique word (UW) of about 20 bits is transmitted, which is used, at the receiver, to establish an accurate time reference in the received burst. It can also be used for the transmitter identification. Al-

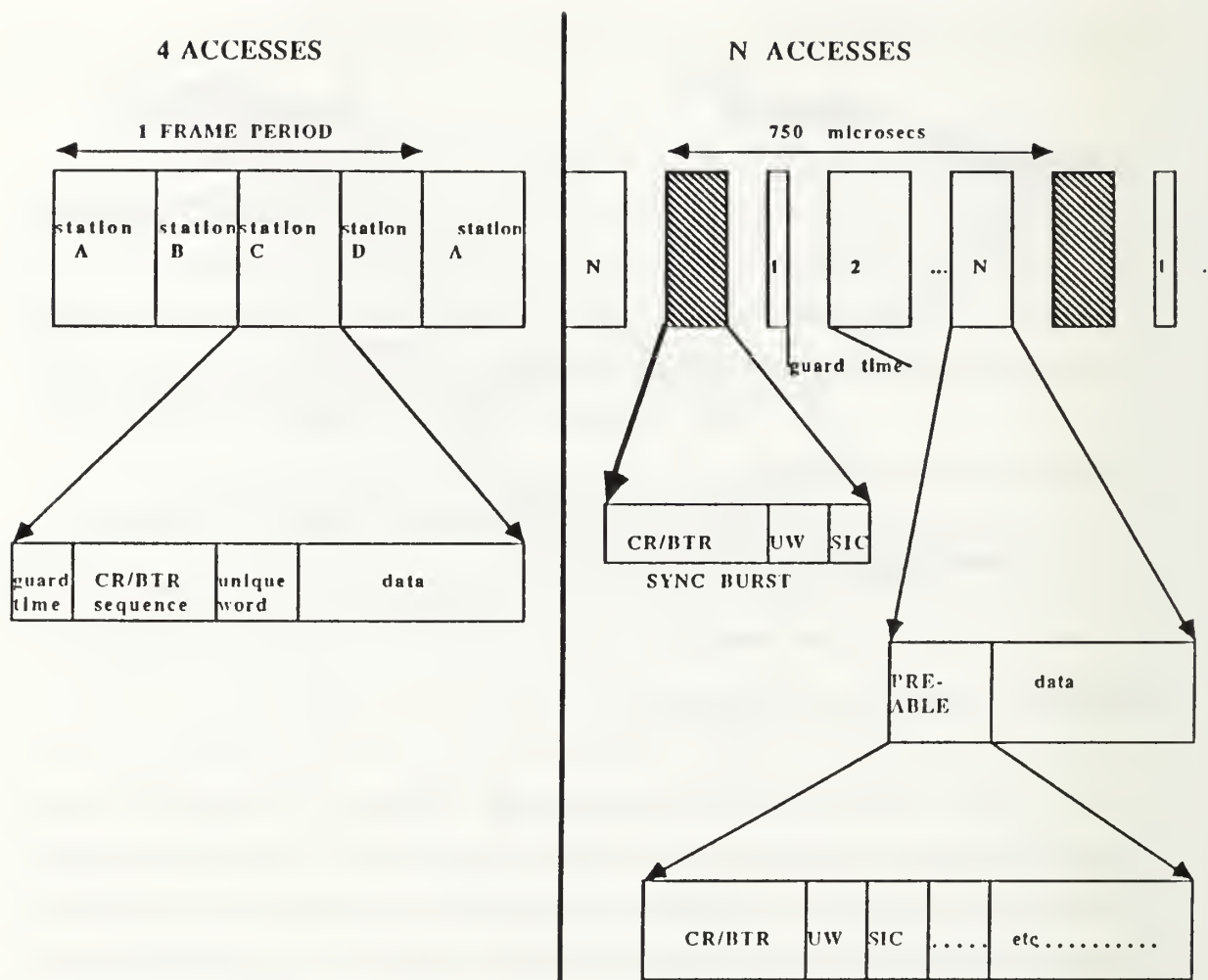


Figure 25. TDMA: frame and burst format.

ternatively, a station identification code (SIC) of about 6 bits can follow the (UW) to provide identification function.

The probability of missing or falsely detecting the UW is of primary importance in practical TDMA systems, for obvious reasons. Generally a probability of miss or false UW detection of (1×10^5) or better is required at the threshold bit-error-rate (i.e. 1×10^4).

The total preamble will usually be 100 to 200 bits per burst frame. Typical frame lengths from 125 msec to 2 msec are practical for high capacity TDMA. Usually the resulting percentage of overhead to total bits per

frame is 5% and the data transmission efficiency is 95%. The total time allocated to guard time is only 0.3% of the total frame time.

(2) *Transmission efficiency - Channel capacity of TDMA systems.* The frame length, preamble length, and number of accesses or bursts per frame determine the transmission efficiency of a TDMA system. The frame length is directly proportional to the efficiency.

Formulating the above, we can get the digitized voice encoded channel capacity as:

$$C = \frac{1}{V} \times [R - \frac{NP}{T}]$$

Where:

R: relay transmission link bit rate

V: bit-rate for one voice channel

P: number of bits in the preamble

N: number of bursts in frame

T: the frame period

Thus, TDMA can theoretically provide the highest channel capacity because, in theory, the only inefficiency arises from the guard times between adjacent slots. However, TDMA requires that analogue signals be digitized and this necessitates analogue-to-digital conversion.

c. Code Division Multiple Access (CDMA)

In CDMA, the transmissions of each earth station are spread over the time-frequency plane by a coded transformation. In this scheme, every accessing station uses the entire repeater bandwidth, to transmit digital data which is impressed on the RF carrier by code modulation. A long code sequence, typically a 2^n P-N one, is transmitted for each digital data 1, and either no code or separate P-N code is transmitted for each data 0. The sequence length and data rate are chosen so that the RF signal spectrum fills the repeater. At the receiving end station a correlating code recognizer is re-

quired. This usually consists of P-N sequence generator and a recognition gate followed by an integrator. When the correct code is received, and the local generator is in phase, a stream of ones will flow from the recognition gate, and reception of a predetermined number of ones is taken as indication of a code-bit (Figure 26). Again the synchronization problem is of vital importance. The receiving station has to synchronize its code generators with the incoming code. The advantages of the method are similar to the ones mentioned for the spread spectrum techniques.

To date, spread spectrum multiple access (SSMA) has been used only in military SATCOM systems, with narrow-band transponders (2-5 MHz). It is very suitable in this application, where mixed size and capacity stations wish to operate in a net. The overall system efficiency is not high, but jamming of the satellite is much more difficult than with FDMA or TDMA systems.

The multiple access capability of SSMA can be understood more easily by consideration of the spectrum of a long P-N sequence. There will be single spectral lines at the frame repetition rate, and the clock rate, and multiples and products of the two. Not all lines will be present for a given sequence, and different P-N sequences with the same clock rate will have different spectral line structures. (The use of an MSRG with 4 feedback taps for the P-N sequence generation has been tested and the results are shown in [Ref. 5: p.18]. The codes can thus be added to produce a combined spectrum which can subsequently be separated into the original two spectra by a pair of matched filters. The code-recognizers can be thought of as matched filters which respond only to their present spectral line pattern; because of the complex nature of the spectrum, the response time of the filter must be long, as it must respond to spectral lines down to the reciprocal of the frame rate.

SSMA coding is very similar to pulse compression techniques used in radar systems. It produces a single, short pulse from the target echo for a long transmitted code sequence. In a typical case the signal might be a 64 bit digital stream, (8 bit PCM with 8 kHz sampling rate) which triggers a 2^n sequence. If the transponder bandwidth is 5MHz a 64 bit sequence could be

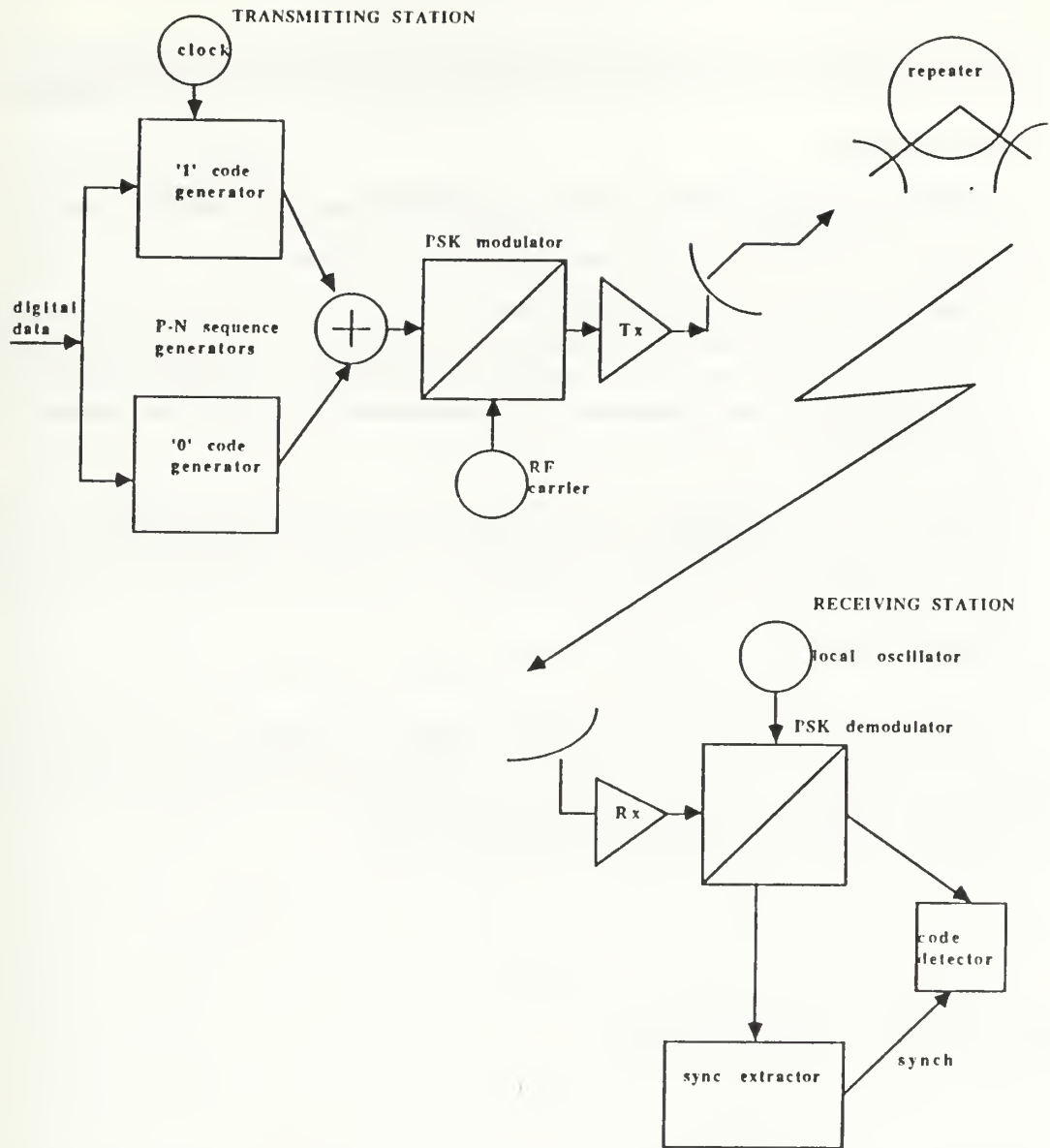


Figure 26. Spread Spectrum Multiple Access (SSMA): PN sequence scheme.

used (2^6), although this would not impart a great deal of security to the system. To obtain security a very long sequence is needed and lower data bit rates must be used. Companding and non-linear sampling in the PCM generator can also reduce the data BER considerably.

III. SPREAD SPECTRUM SYSTEMS IN A JAMMING ENVIRONMENT

A. THE THREAT

A military communications system operating in an electronic warfare environment has to be protected from the following threats:

- Jamming-intended to increase the noise level in the receiver, thus rendering the message unintelligible. This can be done with various techniques-either by 'brute force' through sheer radiated power, or by 'smart jamming' adapting the disturbance to the signal in order to maximize the effect.
- Exploitation-intended to localize and to identify the source; through traffic flow analysis it allows the reconstruction of the command chain in a military operation; through crypto analysis, it gives access to the information in the message.
- Deception-intended to transmit signals formally correct but false.

From a theoretical point of view, the interference that a jamming signal $J(t)$ produces, depends on the amount of its correlation with the desired signal $S(t)$, as expressed by the cross-correlation factor:

$$R(t) = \lim_{\tau \rightarrow -\infty} \int_{-\infty}^{+\infty} J(\tau) \times S(t - \tau) d\tau$$

From a practical point of view, the evaluation of a communication system under jamming conditions is based on the following criteria:

- Jammer Strategy-which options are open to the jammer.
- A theoretical understanding of the performance of the communication system under jamming conditions.
- System performance under laboratory conditions.
- Characteristics of the transmission channel.
- Jammer power and tactics for a variety of geometrical situations.

- Jammer-Transmitter design and its effect on the performance of the platform.
- Complexity and cost of various alternative jamming and communication systems.

1. Jammer Strategy

In any station, the jammer strategy is very much dictated by the communication system parameters and it is a compromise between completely disrupting communications and the limited tactical and technical capability to achieve that, as shown in Figure 27.

2. Operational and Tactical Aspects - Jamming Power Considerations

Airborne or shipborne jammers are the primary practical threats against a naval group operating in the tactical theater. Ground-originated jamming is not neglected but it is situational. It is difficult to compare these two types of threat by considering only the advantages and disadvantages, they appear to have, on a theoretical basis. In a realistic tactical scenario, the maximum range between ships which have to maintain VHF/UHF communications is typically somewhere between 15-30 NM. In this scenario, the threat is much more likely to be from an airborne jammer with its relative unpredictability. Besides, in the case of the shipborne threat, for a variety of obvious tactical reasons relating to its predictability, VHF/UHF communications can be maintained by decreasing the above distance quite dramatically. The theoretical calculations prove that the attenuation factor determines the distance between receiver and jammer for maintaining VHF/UHF communications. This is different in each of the cases of earth and free space propagation. For a free space path the loss is directly proportional to the square of the propagation distance:

$$L_{fsp} = k \times R^2$$

If this expression is referenced to 1 NM it will become:

$$10 \times \log L_{fsp} = 10 \times \log(k \times 1NM) + 20 \times \log(R/1NM)$$

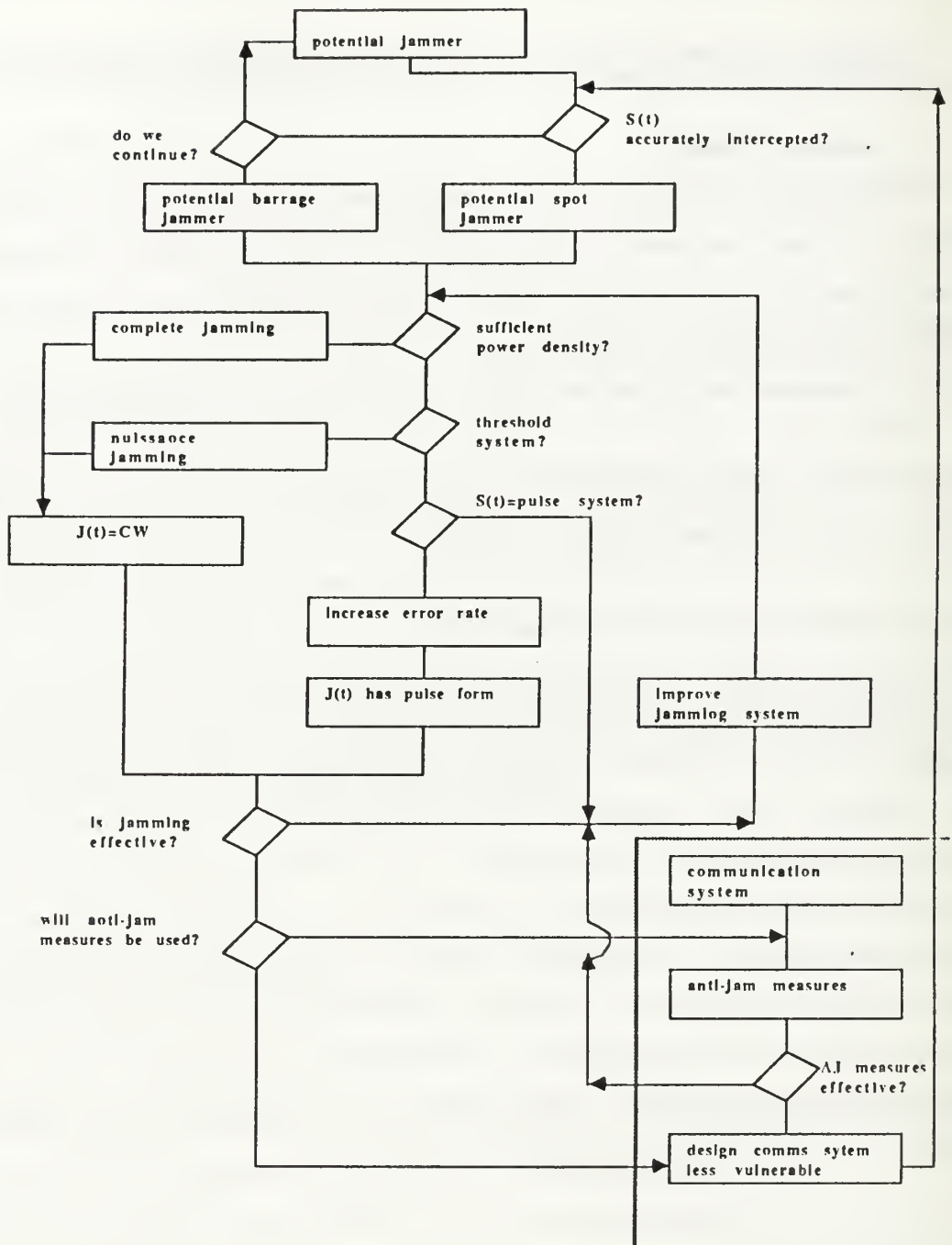


Figure 27. Jamming Strategy: flow chart of a reacting communication system.

For a smooth earth the loss is directly proportional to the distance at the fourth power:

$$L_{se} = k \times R^4$$

or,

$$10 \times \log L_{se} = 10 \times \log(k \times 1NM) + 40 \log(R/1NM)$$

And for a compined sea surface and free space path the loss is:

$$L_{sea} = k \times R^4 \times e^{\alpha \times R}$$

where,

e: is the base of Nepper logarithms

or,

$$10 \times \log L_{sea} = 10 \times \log(k \times 1NM) + 40 \times \log(R/1NM) + 10 \times \alpha \times R \times \log e$$

where,

$$10 \times \alpha \times \log e = \beta$$

$$\beta = 1.71 \text{ dB/NM}$$

Figure 28 shows this result [Ref. 5: p.23].

In the case of a shipborne jammer placed at a distance R_j from the intended receiver, and a transmitter placed at a distance R_t from the receiver, the distance ratio has been found emperically to be related to the jamming power ratio by a fifth power law [Ref. 5: p.23].

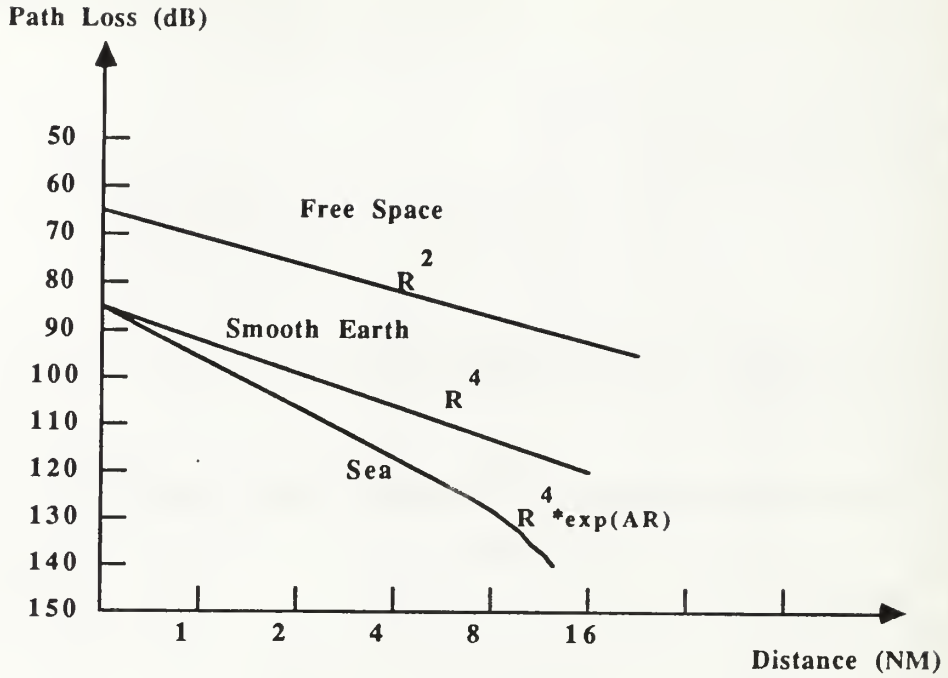


Figure 28. VHF/UHF Propagation: typical curves.

$$\frac{P_{\text{jammer}}}{P_{\text{transmitter}}} = k \times \left(\frac{R_j}{R_t} \right)^5$$

Where,

R_j/R_t : is the distance ratio, and

k : constant factor determining the jamming margin of the particular system, according to the different types of modulation, gain of all antennas involved and the ratio of receiver to jammer bandwidth.

Table 1 provides some results for various distances, from which it can be seen how important the geographical situation is. This is based on standard sea surface signal strength attenuation of 1.7 dB/NM and the fourth power law for a smooth earth: [Ref. 5: p.23].

$$L_{dB} - L_{dB}(1NM) = (1.71dB/NM) \times (R - 1) + 40 \times \log(R/1NM)$$

Table 1. EFFECTIVE POWER OF A SURFACE JAMMER AT VARIOUS DISTANCES FROM THE RECEIVER

DISTANCE (NM)	RELATIVE POWER (dB)
1	0
2	+ 13.7
4	+ 29.1
10	+ 55
20	+ 84.5
54	+ 160

If now the jamming signal comes from an aircraft, the total path loss is a combination of free space plus sea surface attenuation. Besides, there are other factors like receiver-transmitter height and power limitations due to their effect on the airborne platform flexibility that complicate similar calculations. It has been demonstrated that the total path loss for the V/UHF band airborne jamming signal is of the order of 150 dB for distances exceeding 54 NM (100 km). Table 2 gives these results for a frequency of 300 MHz. (Based on a standard distance d_s of 11.6 NM at 300 MHz, a correction factor A as a function of the jammer-to-receiver altitude and the combined formula for sea surface and free space attenuation: $L_{dB} = 10 \times \log k + 40 \times \log R + \beta \times R$) [Ref. 7].

As an assumption, there are two aspects that must be taken into account:

- Complete jamming for all the situations is virtually impossible as it would require infinite jamming power. The jammer can only limit the range of a given communications system.
- The jammer is never certain whether it jams effectively or not.

3. Types of Jamming and their Effects

The most usual jamming waveforms that may be encountered by a spread spectrum system are:

- CW and multitone jammers.

Table 2. CALCULATED PATH ATTENUATION OF AN AIRBORNE JAMMING SIGNAL AT VARIOUS DISTANCES

HORIZON DIS-TANCE d (NM)	HEIGHT (FT)	TOTAL COMBINED ATTEN-UATION (dB)	$20\log \frac{d_s}{d}$ (dB)	A (dB)	TOTAL PATH ATTEN-UATION (dB)
41.2	1500	+ 123	-11	+ 32	144
106	10000	+ 130	-19.2	+ 38	149.5
213	40000	+ 136	-25.3	+ 44	156

- pulse jammers.
- broadband and partial band jammers.
- repeat-back (followers) jammers.

A schematic representation of their frequency spectra is given in Figure 29.

In general, if the interfering signal is a CW tone and appears in the communications channel with an amplitude exceeding a certain level, a sudden and total failure of the data-link system will result. This is due to the "capture effect", peculiar to FM receivers. If the same signal occupies the center frequency, the critical level may be a few dB below the level of the desired signal, because the latter is always shifted to one side of the receiver band. This advantage disappears as the jamming signal drifts away from the center of the channel. Now, in the case of *continuous noise* jamming, (Gaussian-its spectrum is spread over the whole channel bandwidth), a similar threshold effect is also observed, but it is less abrupt on account of the random nature of the noise. The effects of CW or Continuous noise are shown in Figure 30. Statistical results for both these situations are shown in Figure 31.

In the case of digital systems, where pulse transmissions, high bit-error rate, coded and redundant information and economy in mean power are employed for maintaining satisfactory performance, two possibilities of jamming transmissions can be considered, *synchronous* and *non-synchronous*. In the

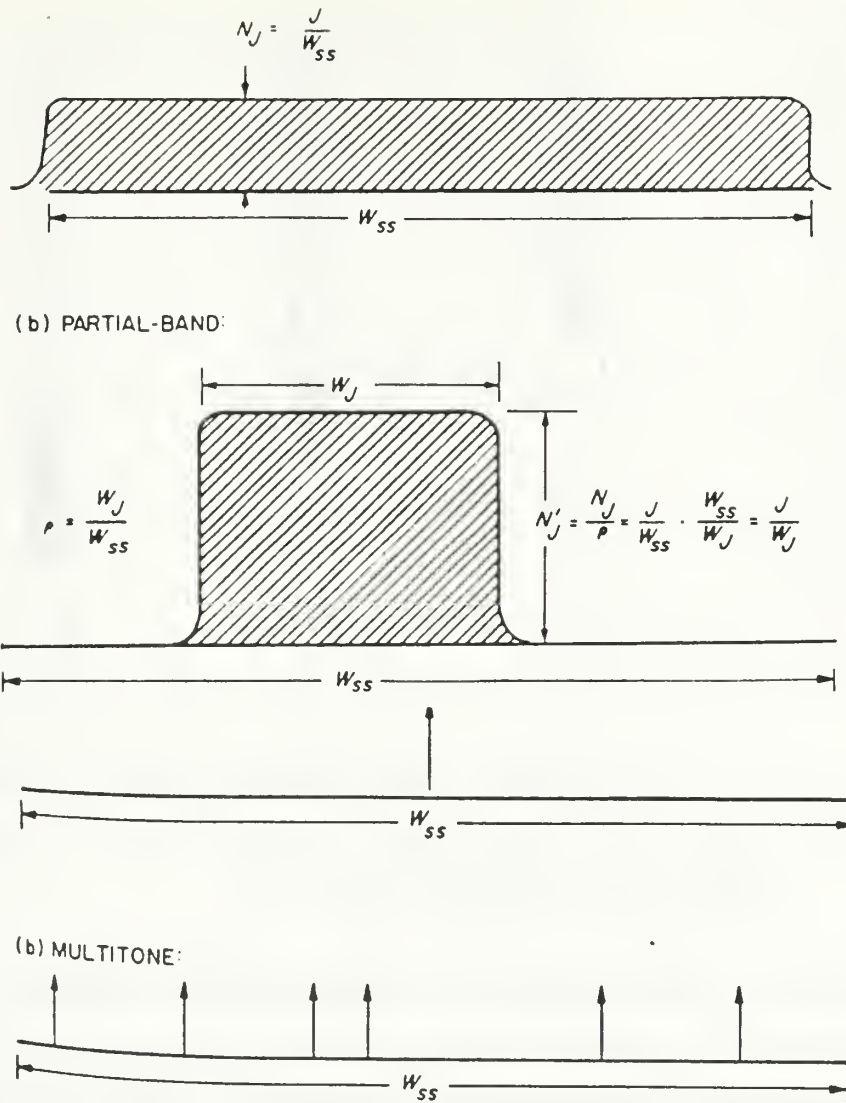


Figure 29. Various Types of Jammers: frequency spectra [Ref. 8: p.140-141].

first case, the jammer's PRF is exactly equal to the message rate, and the jamming pulse occupies the same time slots in the message frame. In the latter one, pulses are transmitted at a rate higher than the message rate, so that at least one jamming pulse falls in every message, but never at the same place. *Synchronous* transmission requires knowledge of the message structure by an enemy. If the data-link transmission can be received at the location of the jammer, it would be quite feasible to use the message synchronization pattern

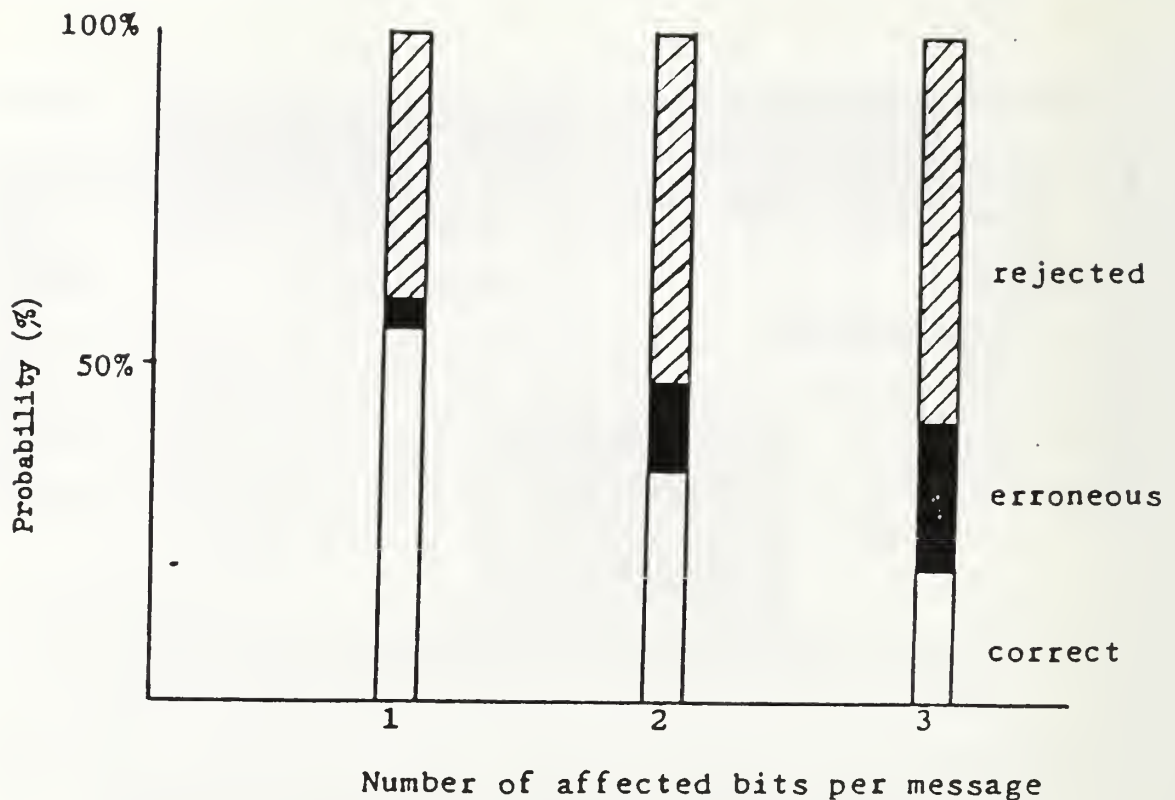


Figure 30. Vulnerability of a Digital Link: synchronous/non-synchronous CW and continuous jamming.

itself to trigger off the transmission of a jamming pulse at a suitable time in the message frame. If at this time a 2-bit pulse of jamming is transmitted on the mark frequency with sufficient power to exceed the threshold, it will certainly cause rejection of the message or reception of erroneous information. Another possibility is a deliberate attempt to mutilate the synchronization pattern of every message. However, this form of jamming is less efficient because a certain amount of distortion could be tolerated by the synchronization circuit at the receiver.

Non-synchronous pulse jamming can be also of CW or noise nature. If the energy transmitted by the jammer is in the form of a pulsed carrier the message error rate will rise proportionally with the amplitude of the jamming pulses, above the critical level, (see Figure 30). The percentage of erroneous

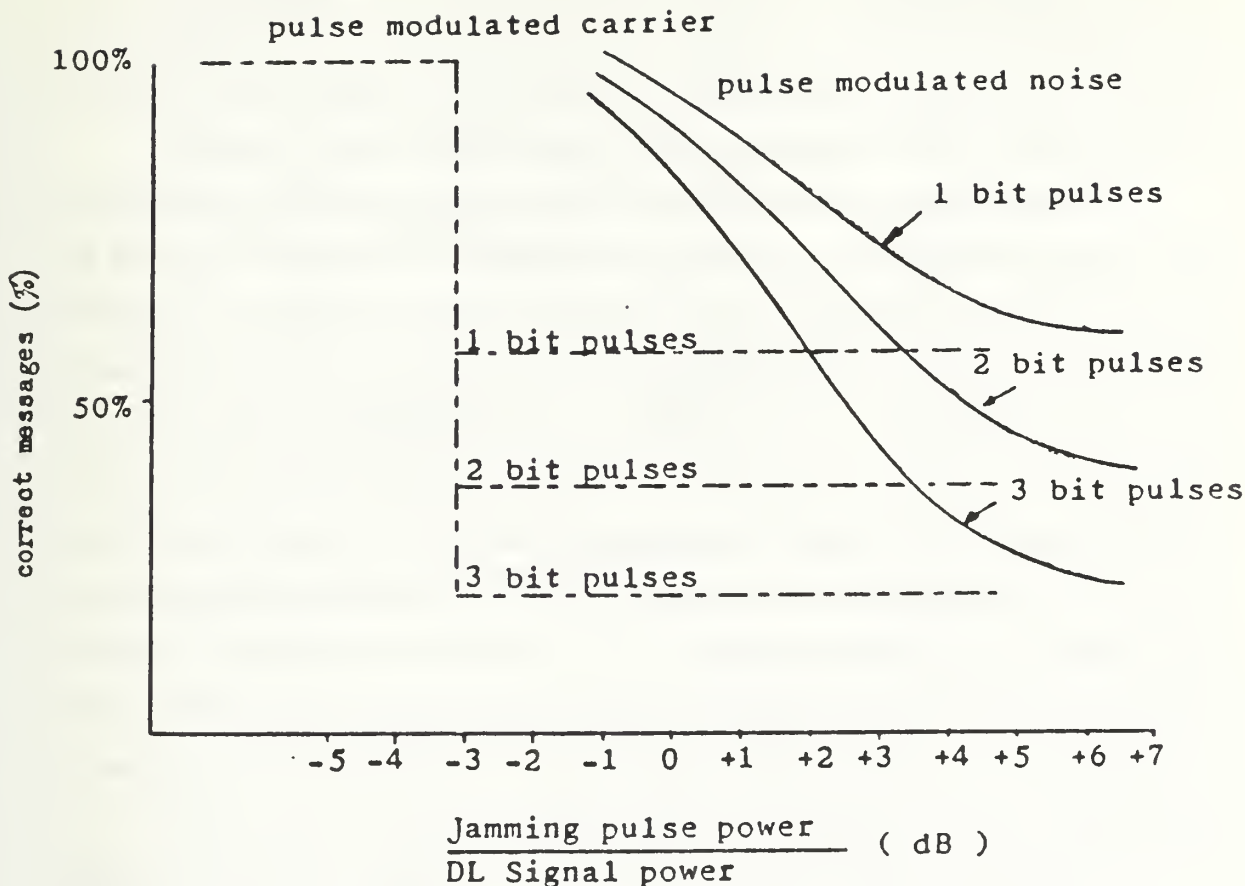


Figure 31. Jamming Effects: results from the introduction of bits of random polarity in the message.

messages would be higher in the case of pulses 2 or 3 bits long recurring at the message rate, than in the case of single bit pulses at 2 or 3 times this rate. If noise pulses are used, the maximum error rates are the same as for a pulsed carrier jammer, but these values are attained more gradually as shown in Figure 30.

This again can be explained by consideration of the probability distribution of the noise envelope. For equal rms values of signal and noise, the percentage of time that the noise envelope exceeds the signal envelope (30 dB above rms) is 10%. In this case the probability that the information bits straddled by the jamming pulse lose their signal polarity is therefore (0.1×0.5)

i.e. 5%. This probability approaches 50% only when the peak power of the noise pulses exceeds the signal power by more than 10 dB (Figure 31). Therefore, whereas CW and noise are about equally efficient when transmitted continuously, CW has certainly a distinct advantage in pulse jamming.

Considering now the situation where the jammer faces difficulties in determining the working frequencies of a data-link system and it cannot set, as usual, the carrier frequency with accuracy, sufficient enough to disrupt the receiver, we come to the conclusion of using *barrage jamming* as a way to overcome all these. This can be either *wide-band jamming*, over a fixed part of the spectrum or *swept frequency jamming*, by scanning the spectrum. Both methods require high power, a considerable amount of which is wasted in unoccupied channels or between channels. In the first case, the power required to maintain a given power density is several orders of magnitude greater than for spot jamming. Figure 32 shows that a large number of transmitters would be needed to cover a part of the V/UHF band (175 MHz wide). A few possible jamming systems are listed below by way of illustration [Ref. 5: p.25].

- 175 transmitters each covering 1MHz, with 10 times more power than spot jammers.
- 18 transmitters each covering 10 MHz, with 100 times more power than spot jammers.
- 3 transmitters each covering 60 MHz, with 600 times more power than spot jammers.

Finally, in the last case, of swept frequency jamming, the scanning rate must be slow enough to ensure that the jamming signal spends a minimum time of, say 400 msec, in the centre part of each channel. The scanning rate is thus given by: $40 \text{ kHz}(\text{freq deviation}) / 400 \text{ msec} = 100 \text{ MHz/sec}$. On the other hand, the repetition rate must at least be equal to the message rate. Therefore the width of the band scanned by one jammer cannot exceed: $100 \text{ MHz/sec} \times 14 \text{ msec} = 1.4 \text{ MHz}$ (14 msec typical data-link message duration). See Figures 33, 34.

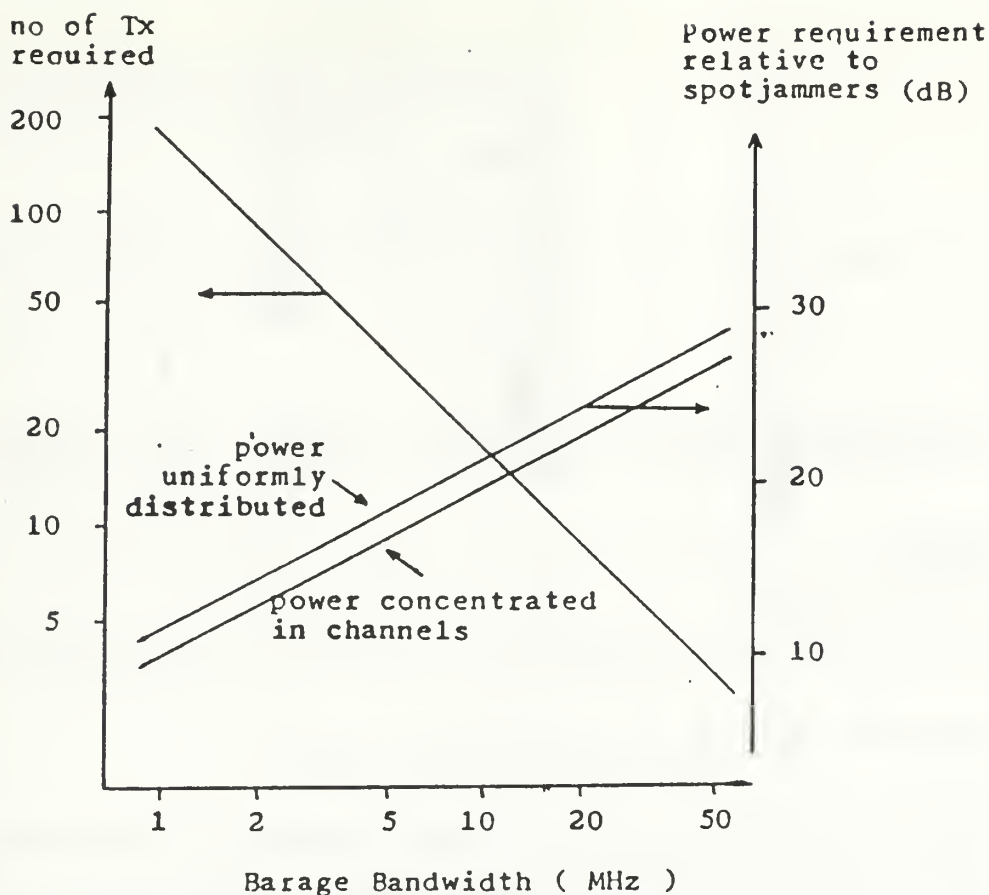


Figure 32. Wideband Jamming: power and equipment requirements.

Thus a total of 125 transmitters would be needed to cover this portion of the V/UHF band. It must be added that in this case the jamming efficiency would be that of non-synchronous pulse transmission.

Another type of ECM would be to make use of gaps in the transmission to trigger off the airborne timing oscillator. This could be done by means of low level synchronization signals. Once the decoding process is started in the receiver, it normally has to be carried out until the end of the cycle. Thus a low powered jammer transmitting short repetitive bursts of false signals (*spoofing by low level signal*), would succeed in keeping the receiver decoders in motion during intervals between messages so that most messages would not be accepted. To reduce the danger of false synchronization, receivers in the NATO

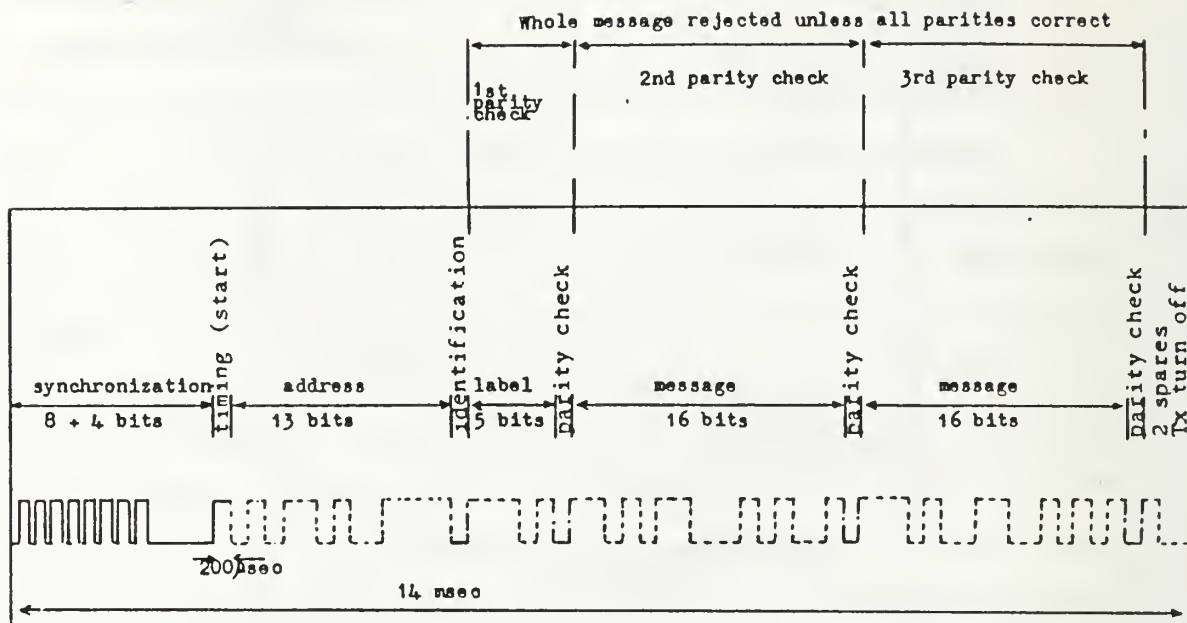


Figure 33. Data Link: message format.

ground-to air link (Link 4A), have been provided with double sync circuits. This procedure ensures that stronger signals are given priority over weaker ones.

B. SPREAD SPECTRUM SYSTEMS PERFORMANCE UNDER JAMMING CONDITIONS

1. Direct Sequence Reaction to Interference

The behavior of a DS system with respect to narrow-band and wide-band interference, and the interference suppression that is achieved in this way is shown clearly in Figure 35, 36. The whole power contained in the spread spectrum signal is compressed in the correlator to a signal bandwidth that is suitable for processing in the subsequent receiver and demodulator. A narrow-band interference signal, such as CW tone, will, on the other hand, be spread by the correlator in the same manner as the required signal in the transmitter, and only a fraction reduced by the value of the spreading factor (processing gain), will fall into the subsequent passband range. In addition to

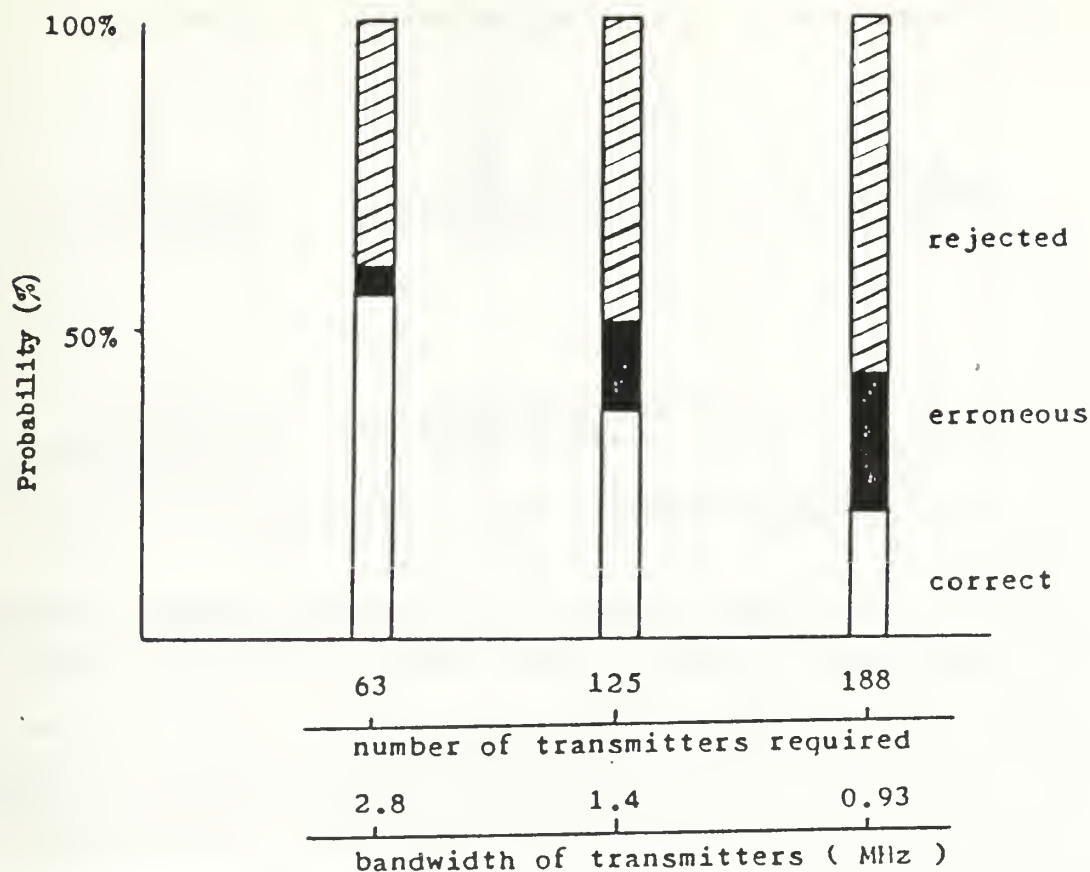


Figure 34. Swept Frequency Jamming: efficiency.

this, the interference signal appears at the output of the correlator as a quasi-noise signal, which means that further processing in the IF circuit will only cause a reduction of the signal-to noise ratio and will not cause a correlated interference.

Other wideband signals that are not correlated with the spreading code, will be further spread in the receiver correlator. This means a far lower power density is present in the processing bandwidth than what would be the case with narrow-band interference. Such wide-band interference can, for instance, be a second DS signal having a different code, and this shows that a large number of DS signals with a differing spread-code can be accommodated in the same frequency spectrum, without intermodulation, adjacent mu-

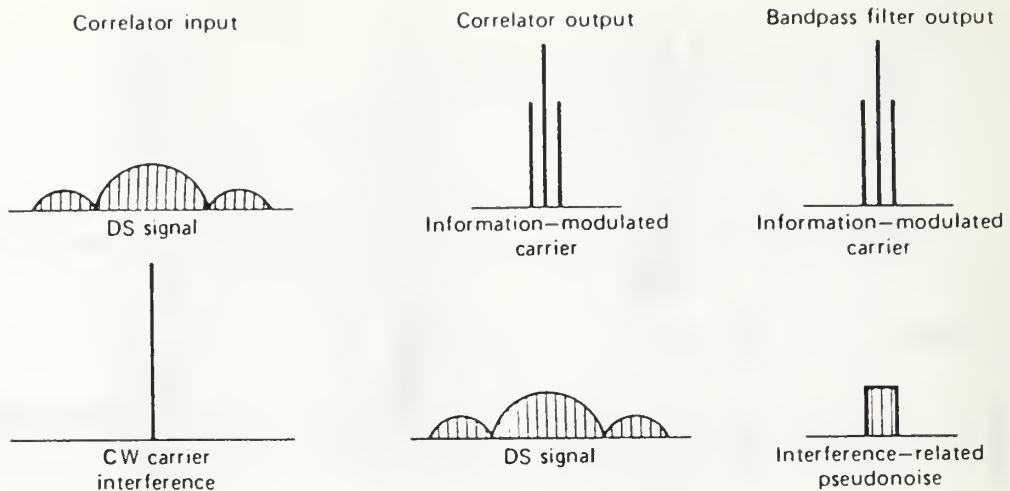


Figure 35. DS System: behavior in the case of jamming [Ref. 4: p.172].

tual interference and safety spacings. But by the time the interfering wide band signal is an almost ideal code sequence with the one used as a code local reference at the receiver correlator, a high cross-correlation can upset the sync detector which would then register no difference between the condition of having interference and having none. Note that the implementation of a cross-correlation integral is a multiplier with two signal inputs (local reference plus incoming signal) and an integrator (LPF). the possibilities of high cross-correlation or no correlation at all are expressed with an output of a maximum value or zero respectively. Thus from above it is assumed that:

- A high amplitude CW tone used at the centre frequency of DS spectrum is the most effective jamming signal than any unsynchronized spread spectrum one [Ref. 9: p.53].
- A high cross-correlated wide-band code sequence would upset the sync pattern, where a low cross-correlated one has no effect at all [Ref. 5: p.27].
- A pseudo noise-modulated jamming signal using the same carrier and chip-rate as the desired direct sequence signal can be most effective, provided that these two elements can be determined by the jammer.

The bit error probability of a DS system, for a continuous and a pulse jammer respectively are given by the following formulas:

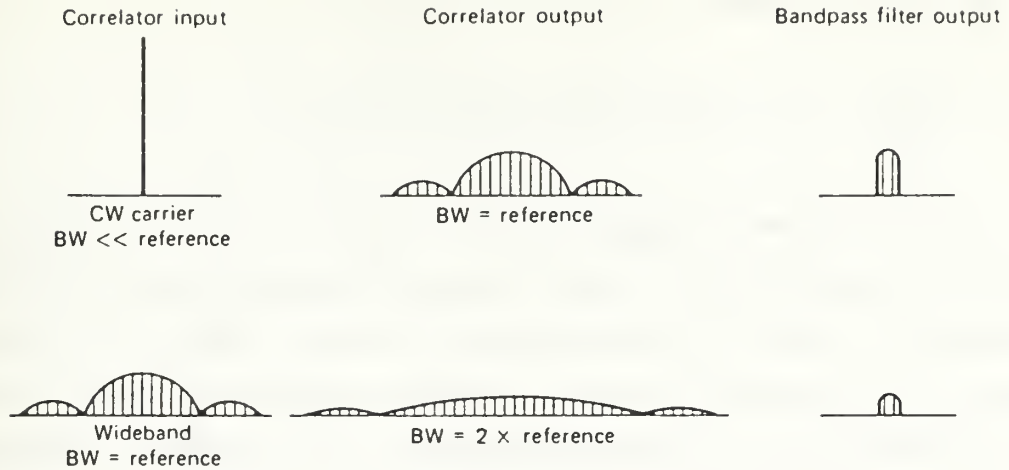


Figure 36. DS System: decorrelation-demodulation [Ref. 4: p.173].

$$P_{BER(BROADBAND-JAMMER)} = Q\left[\sqrt{2 \times \frac{E_b}{N_J}}\right]$$

$$P_{BER(PULSE-JAMMER)} = \rho \times Q\left[\sqrt{2 \times \frac{E_b}{N_J} \times \rho}\right]$$

where,

$$0 \leq \rho \leq 1,$$

$$\frac{E_b}{N_J} = \frac{\frac{W_{ss}}{R_b}}{\frac{J}{S}},$$

J/S: jamming margin,

$$N_J = \frac{J}{W_{ss}}: \text{noise power spectral density,}$$

$$E_b = \frac{S}{R_b}: \text{signal energy per bit,}$$

R_b : information rate,

$$\rho = \frac{W_j}{W_{ss}} \leq 1: \text{fraction of the spectrum jammed}$$

From the formulas above for the two cases of jamming, the first seems to say that regardless of the type of jammer, the performance of DS/BPSK is essentially the same as for the baseline jammer, namely the broadband Gaussian noise jammer. This is contradicted by the pulse jammer example from the second one (r : jammer probability on, $1-r$: jammer probability off) [Ref. 8: p.150].

Figure 37 illustrates the effects of a pulse jammer on a DS/BPSK system. There is a big difference between a constant power jammer ($\rho = 1$), and a worst case pulse jammer ($\rho = \rho^*$), for an uncoded DS/BPSK system. Note that for the same fixed average power J , the jammer can do considerably more harm to an uncoded DS/BPSK system with pulse jamming than with constant power jamming [Ref. 8: p.150].

At any rate, a DS system has a given amount of processing gain G_p . When we try to specify useful jamming margins, it is necessary to know the type of interference, the interference frequency and the amount of cross-correlation between that particular interference and code sequence being used. The performance of the decorrelator, under high jamming conditions for several types of interferers, has been tested for a particular scenario [Ref. 5: p.27]. The results can be quite different for a different scenario, but each time they illustrate the demodulator's difficult task of differentiating between desired and undesired signals.

2. Frequency Hopping Reaction to Interference

As it has already been mentioned, FH is a jammer evasion strategy, rather than one which tries to resist or overcome the jammer. So, terms like jam resistance and jamming-to-signal ratios appear to be irrelevant to FH and enough to be caught by the jammer.

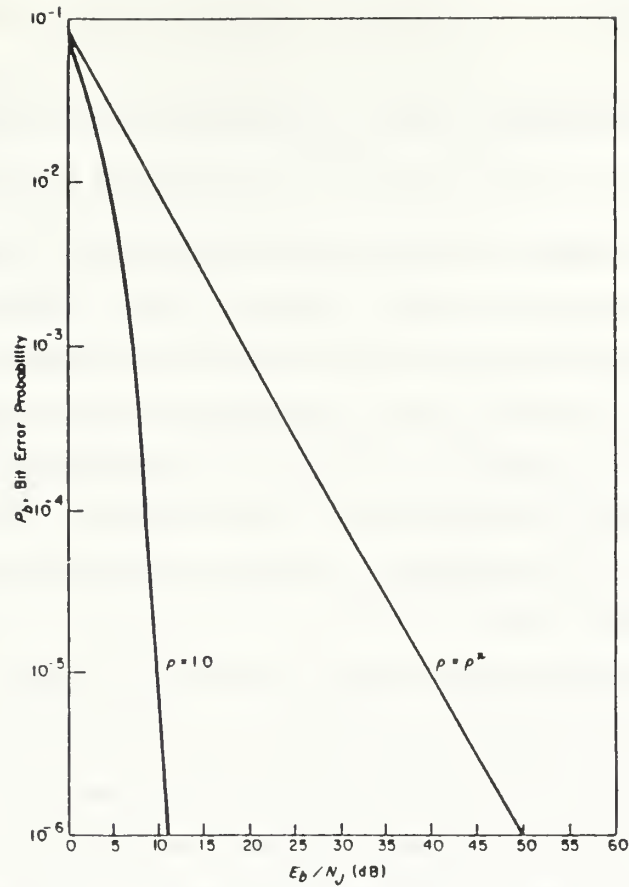


Figure 37. DS/BPSK System: behavior under constant power and worst case pulse jamming [Ref. 8: p.150-151].

Frequency hopping rate and dwell time are the most significant parameters in the design of a FH system, in terms of jamming avoidance. The first one identifies the FH systems as slow, medium and fast. Slow FH is usually thought of as 100 or 200 hps and fast FH as thousands of hops per second (khps), while medium FH is somewhere in between.

But it is dwell time that is regarded as the single most important determination of jammer evasion capability. The shorter the dwell time, the greater is the probability that the system will evade the jammer.

The *narrowband or spot noise jammer* is not a serious threat to any FH system-slow or fast, as it would cause an error of $1/n$, where n is the number

of FH channels. This is met when the jammer-receiver scenario provides the following :

- The jammer frequency has not been deleted from the hopping pattern.
- The jammer has sufficient power to overcome the hopping signal.

Again this would be insignificant if we accept the use of some form of redundancy in the system (multiple chips per bit of information). And even if the jammer can change frequency in an effort to increase the probability of error occurrence, its reaction time is relatively long compared to the dwell time of even the slow hopper. *Barrage jamming (wideband)* has been a serious threat in the past. But the inherent disadvantages of this kind of jammer, such as high power consumption, reduced platform flexibility and mobility, high cost in building, operating and maintaining them plus the fact that they can be easily intercepted and located, made their rise against FH systems almost impossible. Besides when only a few such jammers are deployed against the frequency hoppers, they can be defeated by deleting the jammed band from the hopping pattern. If a large number of barrage jammers are deployed so as to produce band saturation jamming, then the FH system cannot defeat them. But this probability seems more remote to the future, from a practical standpoint.

The most effective jammer against a frequency hopper, would be the one capable of meeting at least one of the following criteria: [Ref. 5: p.28].

- Place sufficient power over the hopping signal power in a large number of channels (*multitone jamming*)
- Be able to break the code and follow the hopping pattern.
- Be able to receive the hopped signal, determine its frequency, and re-transmit a signal calculated to change the effect of the desired signal at the receiver (*follower jammer-transponder*).

Both the last two approaches above can be combatted by effective methods, such as high quality encoding and high hopping rate. *Multitone jamming* appears to be the most effective overall threat to any frequency hopping system. Considering the fact that the effectiveness of the repeater

jammer against FH systems is based on the fast response time, it is assumed that the jamming energy must reach the enemy receiver before the transmitter hops to a new frequency. Therefore, the weighting factors for successful communications are: [Ref. 5: p.29].

- The dwell time of the frequency hopper, (driven by the hopping rate).
- The jammer response time.
- The signal propagation time, (governed by the geometrical situation).

For the transponder to cause any jamming at all, the sum of the propagation time from the transmitter to jammer, the jammer response time and the propagation time from the jammer to receiver, must be less than the propagation time from the transmitter to receiver plus dwell time (see Figure 38).

Since propagation speed is constant, these times can be easily converted to distances ($1 \mu\text{second} = 300\text{m}$) and ellipses can be constructed for each link of the net wherein the transmitters and receivers are the loci of the ellipses. Each ellipse then encloses all points that satisfy the equation described above. Therefore, each ellipse encloses the area in which the jammer must be located to cause any jamming of the target signal. The increase of hopping rate as a mean of combatting a deliberate interferer follows the concept described above. The desired hopping rate is then greater than $(Tr - Td)^{-1}$, where Tr is the total propagation time, and Td is direct path delay. It can be seen then that the minimum required hopping rate is a function of distance to the interfering station and the angle of offset from the direct path (see Figure 38). To give an idea of the magnitude of a possible problem assume that the transmitter and the desired receiver are 10 miles apart and that the interferer is 8 miles from each of them; $Tr - Td = 41 \text{ ms}$ and the hopping rate should be at least 24.2 khps (fast hopping). Then the propagation time protects the system against selective jamming because of the short dwell time. Follower jammers using a set-on receiver have also in the future absolutely no chance to jam efficiently and can only be effective if they intrude into their effective distance (ellipses). But then they are within weapon range. A com-

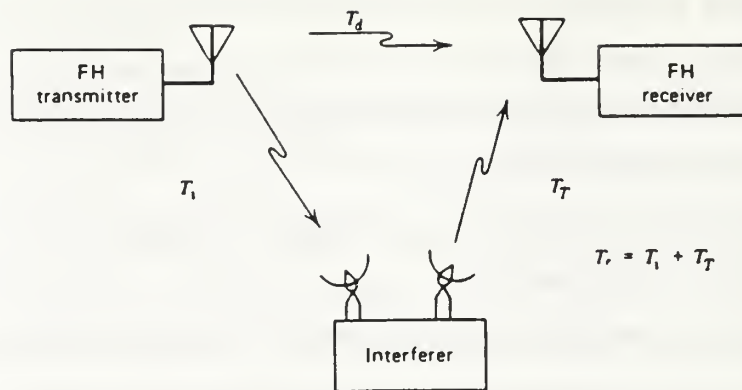


Figure 38. Jamming the FH System: repeater-jammer geometry [Ref. 4: p.39].

parison among a *fast-medium-slow* hopping system is made in terms of jammer's geometrical effects on a link that connects two stations 10 NM apart, and this is shown in Figure 39. With a reaction time of about 20 ms the minor axis of the ellipse is 6 NM and the major axis 12 NM. The second curve shows the effect of 24 khps FH under the same conditions. The 500 hps FH is out of scale. It is a circle with a radius of nearly 200NM. In this example the jammer has to convert more than 30% of a bit to increase the BER. The Fast FH of 16 khps, hops at least once per each bit of information.

There is certainly a trade-off between *fast* and *slow* hoppers in terms of cost, complexity and capability to face the repeater jammer while not allowing it achievable response time. But given the fact that it can follow, then the only way to reduce its effects is by using a non-coherent detector. One approach is to use frequency hopped on-off keying. That is, if a logical 1 is intended, let the transmitter send a signal selected from the library code, if a logical 0, then the transmitter sends nothing. The repeater jammer then has no advantage since transmission of a 1 provides a signal for the jammer to detect and repeat, thus helping the receiver to detect the signal and make the correct decision that a 1 has been sent. When a 0 is intended the desired signal transmitter sends nothing, so the jammer can only make a guess at random. In this case the jammer probability of causing an error is just J/N . But

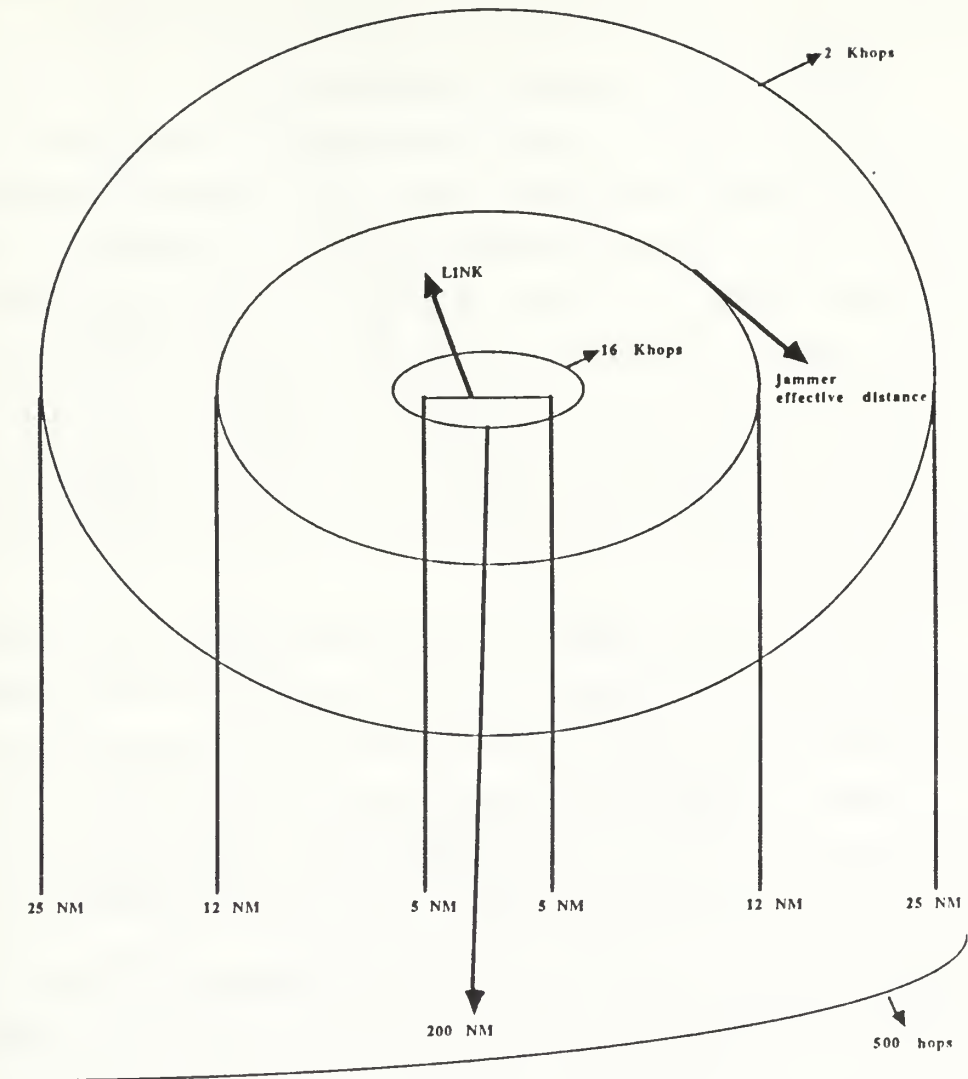


Figure 39. Jamming the FH System: fast-medium-slow hopper geometrical comparison.

when, in the first case, the jammer sends an identical signal as the desired but opposite in phase, an incorrect decision may be caused at the receiver.

Finally, FH systems have their limitations too. The processing gain cannot be extended to infinity, and thus there is an upper limit. This follows that the hopping rate also has an upper limit, as the code rate is limited. Figure

40 below gives an illustration of the effects of several types of jammer on an FH/BFSK system. The detailed analysis is given in [Ref. 8: p.167-181].

3. Hybrid System Interference Performance

If the jamming power is less than the product of the number of hopping channels times the average signal power then it is usually advantageous for the jammer to concentrate its power in part of the hopping band. Assuming a FH/DS hybrid system, when the power of an interfering signal falls wholly within this sector, it is spread by only the bandwidth of the DS signal, whenever the signal hops into this sector. As a consequence when the actual signal lies in this part, a narrowband interferer can cause a loss of performance, with significantly less power than is required to interfere with the entire FH/DS signal structure.

Thus, for hybrid systems, the jamming margin is not related to the processing gain as it has been previously mentioned. It can be significantly less than the simple FH or DS system, although the composite processing gain will be higher. The reason for emphasizing this possible loss in partial-band jamming margin is that when the separated direct sequence channels are used as access channels for multiplexing more than one DS signal on a frequency hopping carrier, it is possible for a single frequency interfering signal to interfere with only one channel, with no significant effect on the others. To prevent one user (or channel) from being completely lost, while spreading the effects of interference across all channels, it is necessary to use each channel periodically, so that each user employs all available frequencies equally. In that case, even if an interfering signal causes complete loss of one channel, each of the multiplexed signals would be affected only a part of the time equal to the ratio:

$$\text{'Number of jammers/Number of channels'}$$

Thus if a 20-channel system had an interfering signal in one channel, it would be affected 1/20th of the time, and if all users shared that channel, a particular user would be affected 1/400th of the time.

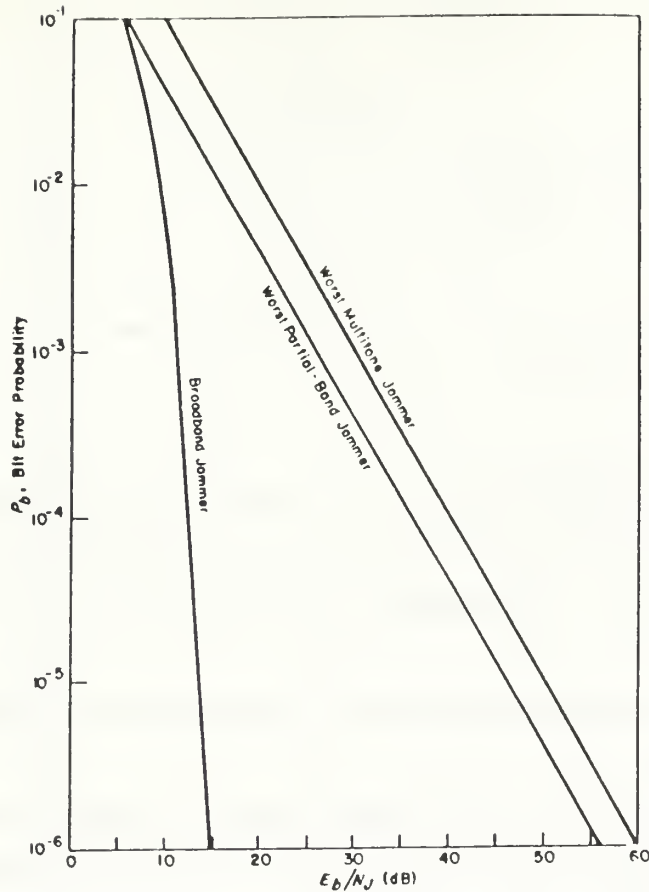


Figure 40. FH/BFSK System: jammer effects [Ref. 8: p.174].

The bit error probability for a FH/QPSK system and for the worst case jammer (WCJ) is given by:

$$P_{BER(WCJ)} = Q\left[\sqrt{2 \times \frac{E_b}{N_j}}\right]$$

For a hybrid FH/DS/QPSK system the respective error probabilities for each of the cases of partial band jamming (PBJ), spread tone jamming which looks like an equivalent AWGN and for the worst case jamming are:

$$P_{BER(PBJ)} = \rho \times Q[\sqrt{(2 \times (\rho \times \frac{E_b}{N_J})^8)}]$$

$$P_{BER(AWGN)} = Q[\sqrt{(2 \times (\rho \times \frac{E_b}{N_J})^8)}]$$

$$P_{BER(WCJ)} = Q[\sqrt{(2 \times (\frac{E_b}{N_J})^8)}]$$

If one compares the $P_{BER(WCJ)}$ for both the FH/QPSK and the hybrid FH/DS/QPSK systems he will conclude that this jammer is less effective against the hybrid system. See Figure 41. A detailed analysis of this is given in [Ref. 10: p.291-295].

4. Near/Far Effect and the Inherent Disadvantage of PN Modulation (Mutual Interference)

The use of Direct Sequence Spread Spectrum offers the unique advantage of LPI (Low Probability of Intercept), as this signal appears as a low power noise and therefore it can be used in covert operations or during periods of radio silence. It has already been mentioned that this noise-like signal is a result of bandwidth spreading which is directly proportional to the processing gain. It has been also mentioned that the processing gain cannot be increased to infinity. Not only because the technical limitations prevent the respective increase of the bandwidth or reduction of the information rate, but also because the receiver can be jammed by a nearby spread-spectrum transmitter, if the latter exceeds the power level plus the process gain of the wanted signal. This can happen even though this transmitter is using a different code. This is called the near/far effect and limits the use of this type of modulation to a link rather than a net. One way to combat this is by the employment of TH/FH hybrid form systems, as has already been mentioned. But this requires exact synchronization for the time and hopping pattern, to avoid cross-interference (cross-talk).

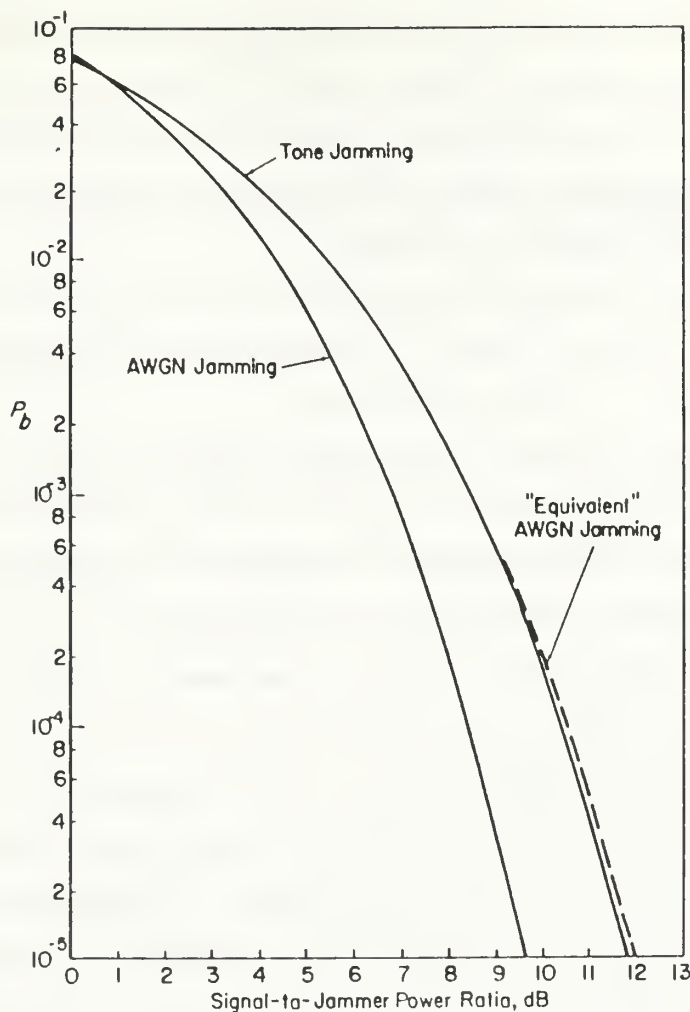


Figure 41. FH/DS/QPSK System: jammer effects [Ref. 10: p.294].

C. TECHNIQUES AGAINST JAMMING/INTERCEPTION

In general, spread spectrum processing offers the most flexible means of providing unwanted signal rejection because of its inherent high processing gain, that does not require any design for rejection of any particular kind of interference and geographic situation. But it does not, however, offer the best processing gain for every tactical situation. There are alternatives such as *antenna null forming techniques*, *electronic cancellation*, and *selective rejection* that can be combined with spread spectrum techniques to produce a compatible system with the advantages of spread spectrum and other signal-to-noise

improving methods. Antenna-cancellation techniques (adaptive arrays) employ the directivity of an antenna to reject a would-be jammer. Such techniques have been used but are not always practical because of positioning restrictions. When geometry is fixed, antenna rejection of interference is limited only by the particular antenna's directivity.

The antenna null forming techniques and their use with the spread spectrum receiver are analyzed in [Ref. 11] where a brief explanation of the basic concept is attempted.

By null forming, a minimum in the pick-up pattern of the antenna can be directed towards a strong unwanted signal as in Figure 42. Some 20-30 db rejection of the unwanted signal is usually achievable. The nulling arrangement can be added to existing communication links to give them some ECCM protection, and high rejection of unwanted signals can be achieved. There are problems, however in making sure that when the wanted signal is stronger than the jammer it is not rejected itself. No protection against intercept or DF can be provided, because even if nulling circuits are employed in transmit as well as receive modes, it is not known where enemy ESM facilities are located. On the other hand, antenna nulling is completely passive in operation, and the enemy can have little idea that it is in use. Depending on how the nulling system is organized, more or less user skill is required, but in all cases more skill is required than for the operation of frequency hopping systems. Although a fully automatic adaptivity is required in today's battlefield with the cost consequently escalating beyond that of frequency hopping, this technique combined with other ECCM techniques renders a fully effective system.

Electronic interference cancellation techniques have been proved more useful although they are also limited to those situations, in which few interfering signals occur, (a separate signal simulator and subtractor is required for each signal), and in which relatively non-complex interfering signals are encountered. For effective interference rejection a duplicate of the jamming signal must be generated through an estimation process. Then this replica is subtracted from the incoming combined interfering and desired signals. The

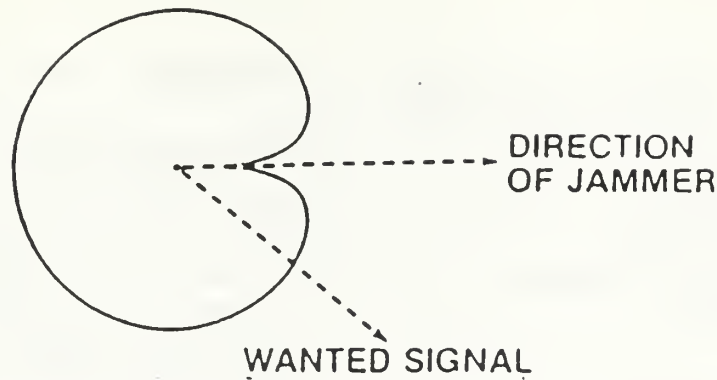


Figure 42. Adaptive Antenna Radiation Pattern: null forming.

result is an interference free signal. Selective rejectors such as notch filters are useful for removing interference signals but are limited to narrow-band signals such as CW. Rejection of a wideband signal with a filter could also reject much of the desired signal, which would degrade system operation and give no gain in performance, even though the interference was rejected.

Finally, burst transmission is an ideal way of transmitting short data messages. It is not suitable for speech because of the relatively large bandwidth and the increased transmitter power required to transmit speech in short bursts. This is a very compact or compressed way of transmitting information, allowing transmission time to be kept short and bandwidth to be minimized. If messages can be kept short, the enemy is obviously faced with a more difficult task in detecting that a transmission is taking place. Manual methods of detection are virtually impossible. Provided that operating frequencies are changed fairly frequently, automatic means of surveillance are also made difficult unless the enemy is prepared to record virtually all the signals in the band he is searching, and then to discard the information that is of no interest to him [Ref. 9: p.60-62].

The processing gains of the various techniques mentioned here are summarized in Table 3 below, for a pictorial comparison [Ref. 4: p.12].

Table 3. COMPARISON OF SEVERAL PROCESSING GAINS PROVIDED FROM SEVERAL TECHNIQUES

SYSTEM	PROCESSING GAIN
Direct Sequence	$(BW_{RF})/(R_{INFO})$
Frequency Hopping	$(BW_{RF})/(BW_{INFO}) = \text{No of channels}$
Time Hopping	1/Transmit Duty Cycle
Chirp	Compression Ratio = rdF
Antenna Rejection	Antenna Gain
Electronic Cancellation	Depends on the accuracy of replica; sometimes 40 dB
Selective Rejection	High; useful only against narrowband interference

IV. SYSTEM DESIGN

The design of a complete Communications, Navigation and Identification (CNI) system through the practical use of spread spectrum technology appears to provide a solution, meeting the requirements expressed in chapter one. These requirements can be translated into specifications for a fully integrated anti-jam tactical communication system.

The development of such a system has been a long lasting effort of the US Department of Defense for all the services involved in joint tactical operations. The product of this effort is currently known as the Joint Tactical Information Distribution System (JTIDS), which features pseudorandom frequency hopping spread spectrum techniques with a low duty cycle signal structure. Originating from the earlier USAF and USN integrated CNI-AJ communication programs, it is being evolved to rationalize the link diversity within NATO. Under this joint effort the US Navy has the leading role in developing the Distributed Time Division Multiple Access (DTDMA) technology. By making use of all the advantages that the spread spectrum techniques offer, the system provides high levels of anti-jam performance and security for both digital data and voice communications, in both the TDMA and DTDMA modes. It also provides precision ranging, relative navigation and conventional TACAN functions. The system provides for total system connectivity.

A tri-service, multi-function, multi-channel system such as JTIDS-TDMA (DTDMA) is the decisive answer to a system architecture suitable to meet the requirements derived by today's complex operational tactical environment. For this reason, a JTIDS-based design will be pursued throughout this chapter. However, certain specifications may be significantly different for smaller-scale requirements such as those which are suitable in closed-sea operational areas. Although reduced capacity, power and EMP hardening levels than the JTIDS baseline standards may be sufficient, cost-effectiveness reasons render

a totally new engineering effort at least uneconomic for the time being, as this is further explained in the last chapter.

An analysis of a JTIDS-based architectural concept and its operational characteristics and functions follows next. The concept of DTDMA is then examined in detail and a description of the system design is given with block diagrams and pictorial representations. The specifications of the JTIDS-TDMA (DTDMA) terminals, the system range calculations and the jamming performance tests are attempted in the following sections. Some of the system deployment considerations are expressed in section 4. Finally the system survivability issues are given in section 6.

A. SYSTEM ARCHITECTURAL DESIGN CHARACTERISTICS

1. Choice of frequency operating band.

A JTIDS-based design operates in the Lx portion of the frequency spectrum (960-1215 MHz), with the frequencies of 1030 and 1090 blocked out so as not to interfere with IFF. The main reasons behind this selection are: [Ref. 12: s.29]

- the favorable line of sight propagation characteristics which provide operation up to 300 nautical miles and beyond from high altitude aircraft.
- the existence of important C-N-I functions such as TACAN and IFF. Particularly, the narrowband TACAN unit included in the JTIDS class 2 terminal shares the antennas with the JTIDS TDMA (DTDMA) signals. But because of the frequency hopping and spread spectrum techniques employed in the JTIDS TDMA (DTDMA) signals, the TACAN and JTIDS TDMA (DTDMA) communications can function simultaneously.
- the low utilization of this portion of the spectrum relative to the lower congested part of the UHF band makes it very attractive for use.
- the ample bandwidth offered by this portion of spectrum (255MHz).
- the existence of large numbers of subscribers among broad classes of users.
- the use of already existing hardware for several functions within this band by a broad class of subscribers supports strongly a time-shared C-N-I system operating in this portion of the spectrum.

2. System functions.

A JTIDS-based design provides integrated anti-jam communications, navigation and identification functions (AJ and ICNI), for tactical combat environment. More specifically:

a. Security and AJ capability

AJ capability is obtained through the use of pseudorandom, fast frequency hopping signal processing techniques. These distribute the transmitted data over several hundred Mhz band. Security is obtained through the use of forward error correction coding (Reed/Solomon) which permits reconstruction of the information content of messages in which up to 50 percent of the pulses cannot be interpreted by the receiver.

b. Communications functions

The system provides information distribution of such critical data as position information on friendly forces, track information on hostiles, threat warning and control and vectoring information. It also provides digital voice as supplementary capability and a common frame reference for the relative navigation function (grid lock).

c. Navigation functions

Precise relative navigation and positioning capabilities are also provided, based on accurate synchronization to a common system of time and a signal structure that permits reliable time-of-arrival (TOA) measurements. The TOA-derived position measurements depend on the source accuracy of the position identification provided by the transmitting terminals, whether it be from a highly accurate satellite Global Positioning System (GPS), an Inertial Navigation System (INS) or a fixed geographic position. Especially in the case of a Battle Group and in blue water conditions (out of radar range of land), it is provided from the position of the flagship. This TOA-derived position estimate is then compared in the JTIDS terminal with other navigation and position information sources to provide improved position estimates (up to a 100 feet). This provides a constant frame of reference for the entire group. Practi-

cally, it eliminates the grid lock and dual track errors currently experienced using Link 11.

Emulation of the TACAN interrogator function in all tactical aircraft terminals is also provided.

d. Identification functions

It provides identification of all system platforms through position and identification data interchange among them and emulation of all the IFF transponder functions in all system command and tactical platform terminals.

3. TDMA Multi-net operation

Figure 43 shows a typical participation of forces in the JTIDS communications ring. These forces can be simultaneously serviced based on the principle of time sharing the same randomly hopped frequencies as other subscribers within the communications net. To accomplish this a time "cycle" (or epoch) is established in which time slots are repeated every 12.8 minutes. The epoch is divided into 98,304 individual time slots of 7.8125 milliseconds each, thus providing 128 time slots per second for the transmission or reception of data. The time slot structure is shown in Figure 44.

Each subscriber (terminal) is assigned selected time slots for transmission based on the particular message requirements necessary to support the mission. It will automatically be able to receive in all time slots that it has not been assigned to transmit. It may have many transmit time slot assignments, consecutive or spaced, within the epoch. Furthermore, a subscriber may share other nets by being programmed to transmit or receive in time slots of numerous other nets, each net functioning in the same time reference and line-of-sight relayed area but at different frequency hopping schemes. Proper synchronization is achieved by designating one of the subscribers to serve as the time reference, by adjusting the crystal oscillator in the various subscriber terminals to the same time throughout the inter-operating nets. Each subscriber can either broadcast or receive on only one net at a time in any particular time slot. Typically, more than one subscriber would not be assigned to transmit in the same time slot on the same net, unless it was beyond the others

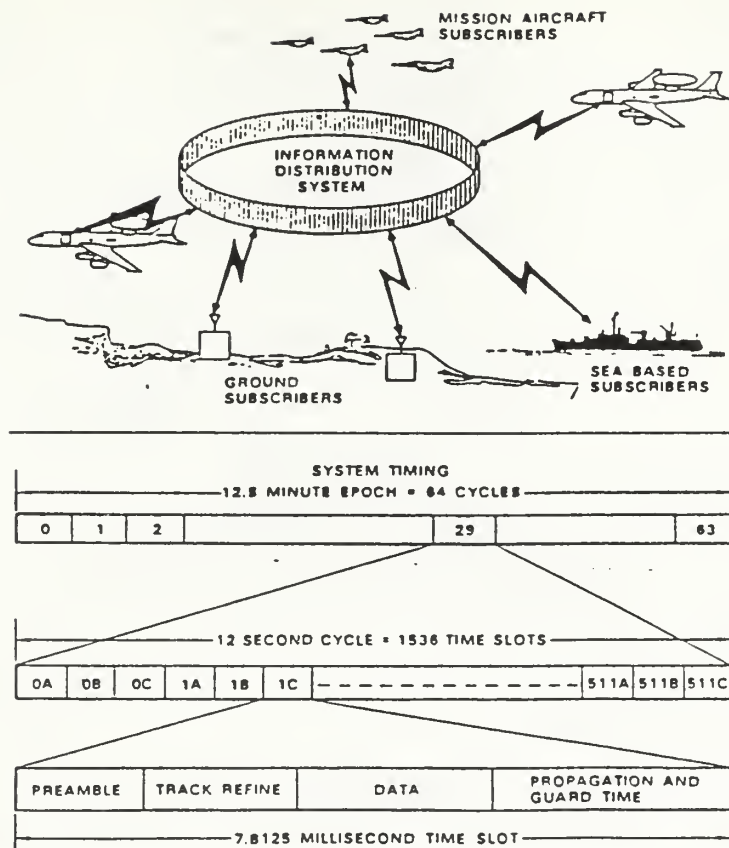


Figure 43. JTIDS Communications Ring: typical participation of forces [Ref. 4: p.329].

required operational range. In the case where a receiving terminal may be within the propagation range of more than one terminal transmitting in the same time slot and net, then the receiving terminal will process only the first signal received, (being from the closest to it transmitting terminal). Any number of subscribers can receive in the same time slot of the same net, thus realizing a one-to-one, a group-call, or an all-call capability in selected time slots. Moreover, subscribers can receive information from time slots while not transmitting in any slot. This feature permits a subscriber to function in a radio silent mode while still receiving mission and threat information over JTIDS nets.

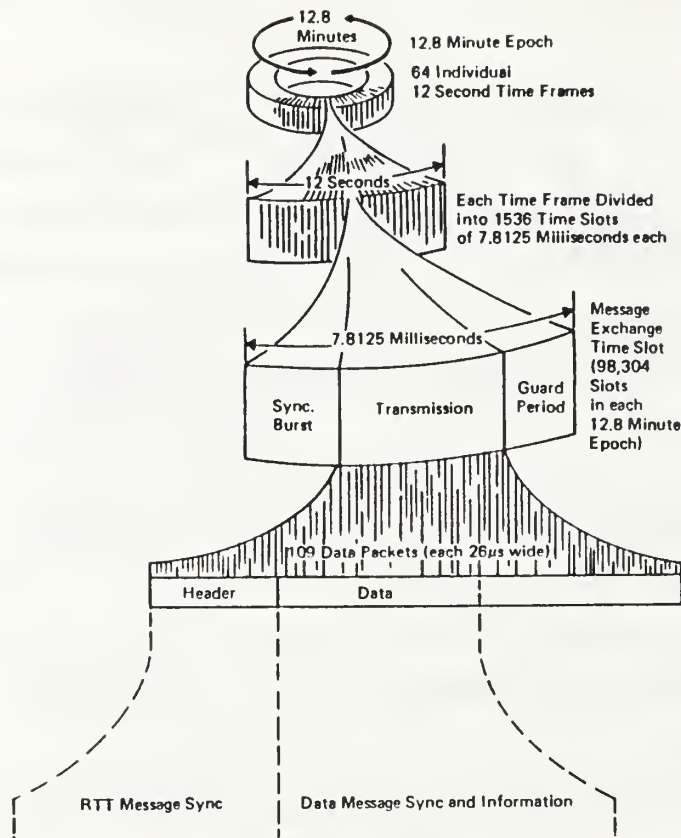


Figure 44. JTIDS TDMA: time slot structure [Ref. 13: p.35-11].

In summary, every subscriber can broadcast data in a commonly accessible communications data stream, represented by the ring in Figure 43, above. All other subscribers can receive information by continuously monitoring and sampling the database. Since this is a broadcast system, every subscriber after deciding what category of data he wants, he can receive all the information the system has at this particular period in that category [Ref. 14: s.26].

4. Message Standards and System capacity.

A particular message broadcasted into the database may be formatted or unformatted. The former are intended to replace much of the traffic carried on UHF voice channels where the latter may accommodate digital voice messages.

Perhaps the most applicable message standard for a system like this, is one that supports both data and voice, relative navigation, identification and any other information than can be digitized. A JTIDS-based design provides three basic message structures with different information capacities which can be matched to the type of information being transmitted. These are shown in Figure 45. The standard double-pulse structure is the most rugged from a performance standpoint. Other structures permit the packing of two and four messages in the time slot through the use of the single-pulse structure and the deletion of the pseudorandom message start (jitter).

Several processing steps are followed before the system digital information signal is transformed into a reliable and jam-resistant form for transmission. The net effect of all these steps is a signal that is spread across a wider spectrum and is very robust to error detection and correction. The basic system message corresponds to a time slot of a 7.8125 msec duration. It consists of a block called 'jitter' which is used to vary the actual start of the data transmission in each time slot. The next two blocks correspond to the sync and timing data, that determine the transmission of the message. Then the message block follows next and the final block is the guard one that allows for the propagation of the transmitted signal to a range better than 300 nautical miles without other transmissions causing interference. The guard block has a duration of 4.4585 msecs.

The message block starts with 210 bits. During the first processing step a polynomial encoding is performed on these bits. As a result, the message becomes 225 bits, with 15 parity bits added for error detection. In the next step, the data is spread pseudorandomly with direct sequence modulation. This is done by mixing 5 bits at a time with a specific sequence of 32 bits (chips), depending on the pattern of these 5 bits. The sequence has a frequency of 5 Mhz and it lasts for 200 nsecs. This is called cyclic code shift keying (CCSK) and it results in changing each 5-bit group of data to a 32-chip symbol of 6.4 μ secs duration. This coincides with the hop transmission time and so each symbol is ultimately transmitted over a different carrier fre-

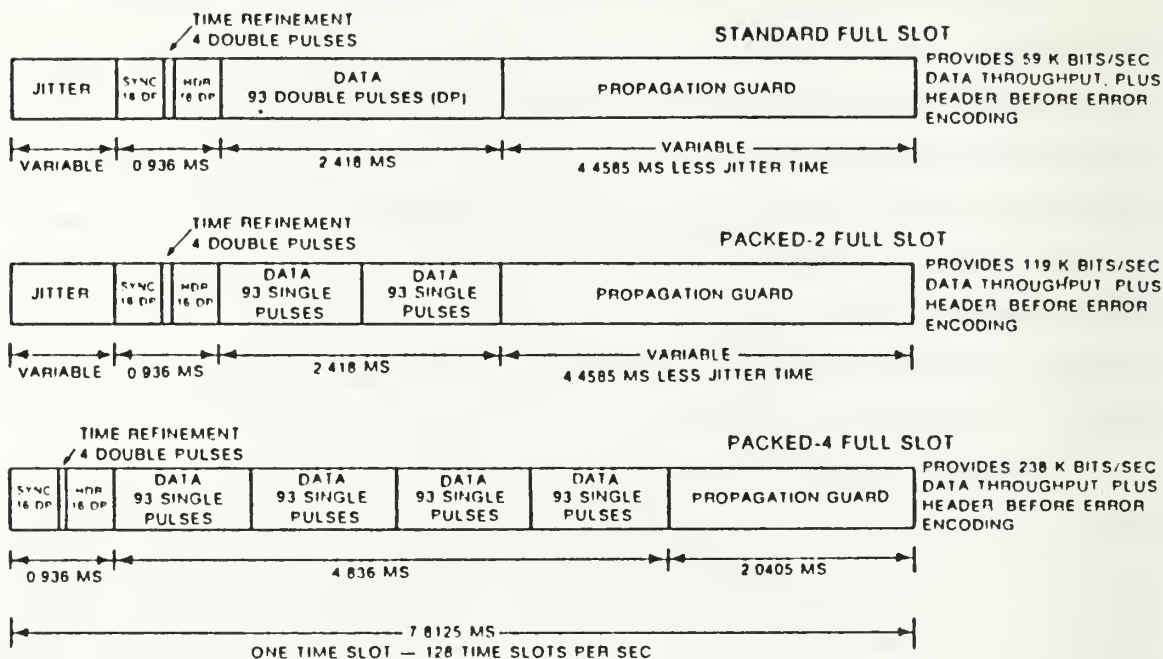


Figure 45. JTIDS TDMA Message Structures: slot variations and throughput capacities [Ref. 15: p.9].

quency. Another result of this step is that each 32 chip sequence has a Hamming distance of 16 bits, which means that 16 erroneous bits can occur before the symbol can be no longer restored [Ref. 16: s.30-3]. The net result of this step is 45 symbols, each representing 5 bits of the original data. The Reed/Solomon error correction code is applied to the symbols, by taking 15 symbols at a time and adding 16 overhead to create a 31 symbol codeword, that can be corrected if up to 16 of the symbols are changed or lost. The result of this is 3 codewords of 31 symbols. Then each of these 93 symbols continuously phase shift modulates (CPSM) a different carrier frequency consistent with the hopping scheme, before transmission. This particular time of RF modulation provides very little in-band and out of band interference as it gives a spectral rolloff of $1/f^4$ [Ref. 16: s.30-2]. The net effect of these steps is that the message now contains over 100 percent overhead data. Since each message contains 210 bits of useful information and it lasts for 7.8125 msecs, it follows

that the system throughput is essentially 26.880 Kbps. Table 4 shows the characteristics concerning the system standards.

In the standard message transmission scheme, double pulses are employed, whereby two pulses carry the same information (at different frequencies) for more jam resistance, thus the applied data rate is half the maximum possible. When forward-error correction is applied, the data throughput is reduced by as much as 50 percent more. The Packed-2 and Packed-4 time slot structure can be employed to increase throughput capability. The first one employs the same elements as in the standard structure. However, in Packed-2 redundant information is not transmitted on every two pulses, thus doubling the data throughput capability over the standard one. In the Packed-4 structure, the "jitter" time is eliminated and the silent period is reduced, thus doubling the number of data pulses that can be transmitted in the time slot. Also, like the Packed-2 structure, redundant information is not transmitted on every two pulses. So the data throughput here is four times that of the standard structure, with forward-error correction applied.

5. System modes of operation.

The JTIDS is designed to operate in four communication modes, giving the user the choices of narrow or wide band, secure or non-secure transmissions. In mode 1, the signal is secure and provides maximum jam-resistance through frequency hopping (through its 255 MHz band excluding 1030 and 1090 IFF frequencies). This is a shared band with TACAN frequencies (up to 128 multiple nets may be created in this mode, limited only by self jamming when multiple nets are used in the same geographic area) and pseudorandom noise. This mode provides the most security in a hostile EW environment. Mode 2 uses a total bandwidth of 10MHz without frequency hopping. Mode 3 uses identical synchronization in all time slots. And in Mode 4, neither the signal nor the message is encrypted.

Table 4. SYSTEM TECHNICAL CHARACTERISTICS

INFORMATION FORM	
Bit oriented messages	
Digitized voice	Continuously Variable Slope Delta
RADIO FREQUENCY SPECTRUM	
L-band allocation	960 to 1215 Mhz, 153 Mhz band-width
Frequency hopping channels	51 frequencies spread 3 Mhz apart
RADIO FREQUENCY PULSE	
Center frequency	hopped over 51 frequencies
Duration	6.4 μ secs
Bandwidth	3 Mhz
Chips per pulse	32
Chip rate	5 Mhz
Chip modulation	CPSM
SYMBOL ENCODING	
Pulses per symbol	1 or 2
Bits per pulse	5 bits
Chips per 5 bits	32
Error detection/correction	Reed/Solomon (31,15) character code
Direct sequence spreading	cyclic code 32 chips per 5 bits
MULTIPLEXING SCHEME	
TDMA or DTDMA	128 slots/sec (7.8125 msec/slot)

6. DTDMA (vs) TDMA architecture.

A JTIDS-based design uses both TDMA and DTDMA techniques. They both use the same basic JTIDS pulse shown in Figure 46. Their main common characteristics are:

- Both employ the same signal pulse (6.4 *musec*), pseudorandom modulation of the individual signal pulses and fast frequency hopping on a pulse-by-pulse basis.

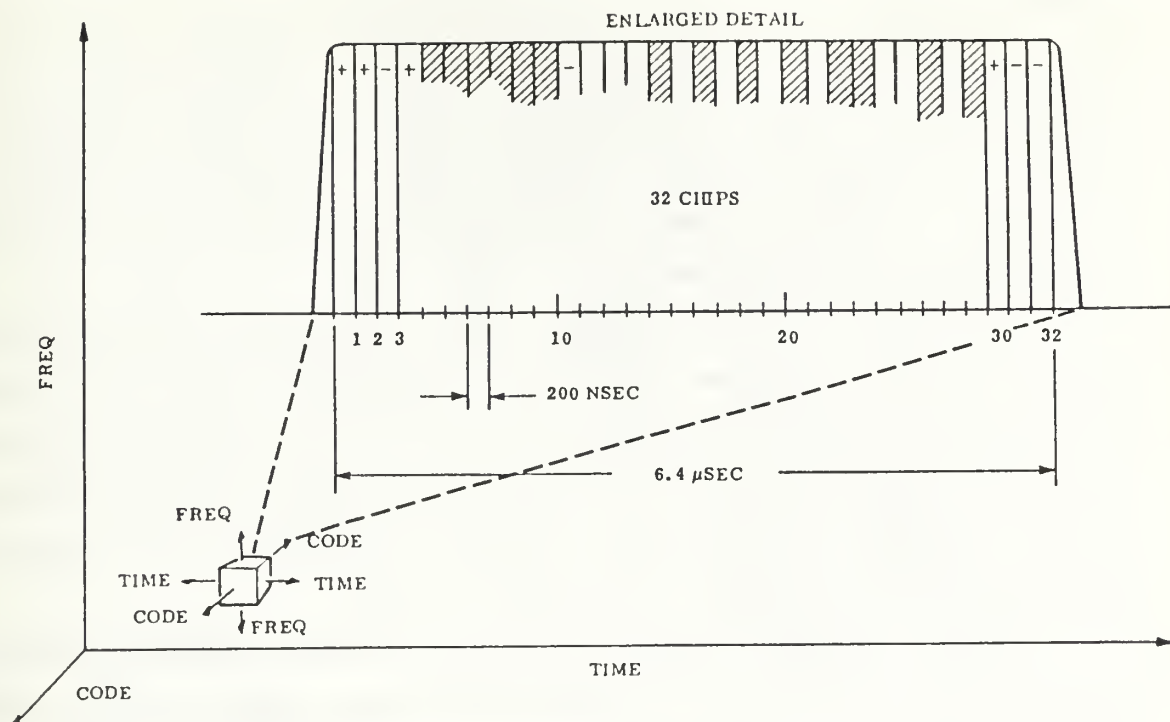


Figure 46. JTIDS TDMA/DTDMA: basic pulse [Ref. 17: p.37-6].

- Both provide a secure anti-jam link for data and voice.
- Both use the same crypto (secure) data unit.
- Both operate in the same RF spectrum (960-1215 MHz).
- Both employ the same Reed-Solomon forward error correction code.
- Both can do relative navigation.

The main difference between the two techniques is the way time is used with each one. JTIDS-TDMA has already been analyzed in the previous sections. In DTDMA, the low duty cycle pulses are pseudorandomly distributed in the time-frequency-code (T-F-C) domain. This is shown in Figure 47. The pulses sync data here for a given message are interleaved in a pseudorandom pattern with pulses from different messages. In TDMA, the pulses constituting a particular message are sent together. This extra characteristic of pseudorandom time separation and higher capability in DTDMA is the principal feature that distinguishes it from conventional TDMA. This ar-

chitecture provides the flexibility of tailoring the communication requirements of a particular force to its composition needs. As messages are not constrained to any time slot or fixed time duration, they use only the required portion of the system capacity, permitting system resources to be used and reallocated in the most efficient manner. This way, high levels of AJ and LPE performance are practically realized.

The DTDMA channelization and signal structure are based on maximizing the utility of available ambiguities in time frequency and phase codes. If all of the LX-band (T/F) space is subdivided into elemental resolution elements, then they may be formed into specific RF channels which may be addressed by particular pseudorandom patterns. These then may be used for specific functions by selected terminals [Ref. 5: p.20].

DTDMA channel architecture is modular in nature. A general depiction of this is given in Figure 48. It consists of four basic levels. The basic event (BE), the basic channel (BC), the function channel (FC) and the metachannel (MC). The fundamental building block in the T-F-C space is called the basic event. The pulse transmitted during the BE has a duration similar to the basic JTIDS-TDMA pulse of $12.8 \mu\text{secs}$. The BE is uniquely associated with an event code that fully describes its signal and channel parameters, such as carrier frequency, phase code, transmission time e.t.c. These event codes are generated at a rate of $12.8 \mu\text{s}$ (78,125 times per second). The BC is the pseudorandom addressing of one only event per page in the T-F-C domain, where a "page" is defined as the time duration needed for 512 BE's or 6.55ms ($512 \times 12.8 \mu\text{s}$). Each BC contains essentially one BE per page, thereby dividing the page over the BC's for transmission.

A contiguous stream of pages forms by definition a metachannel, which is essentially the totality of all basic channels on any net ($1 \text{ MC} = 512 \text{ BCs}$). The MC may be viewed as the multiplexing of many T-F-C patterns each with mutually exclusive BEs. A MC has a total of 78,125 different BEs which constitute 512 BCs, and a BC event rate is 152.2 events per second, ($78,125/512$). Different T-F-C patterns define different MCs. A function channel (FC) now is the to-

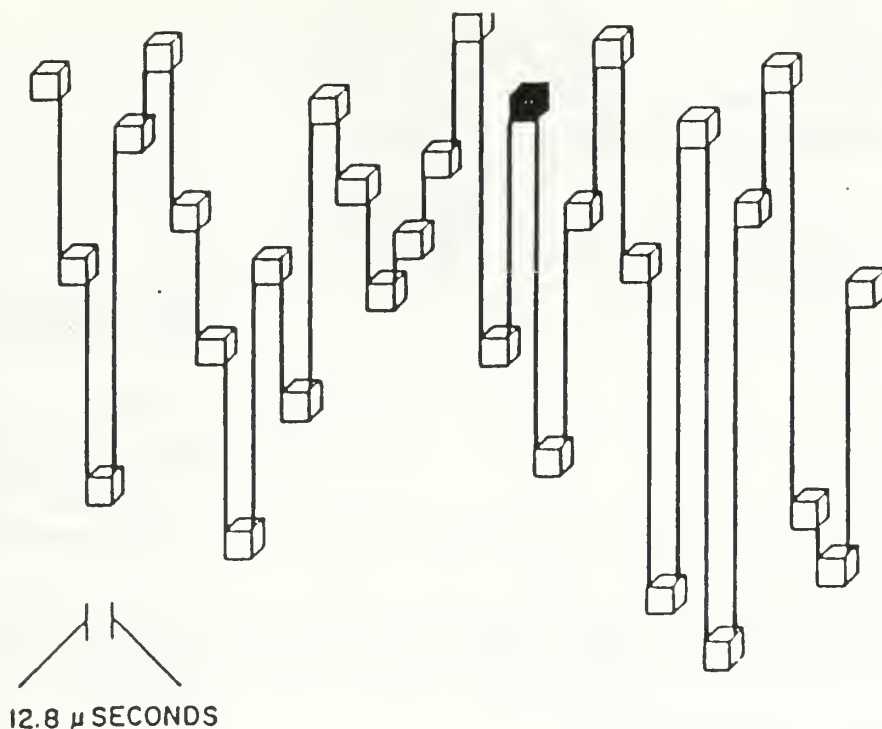


Figure 47. JTIDS-DTDMA: T-F-C space representation.

tality of building block channels, which are sets of BCs organized to support the required numbers of subscribers with a prescribed level of channel capacity. The specific user functions, which such channels are designed to support, can include a command and control intercept link like Link-4A, a surveillance reporting link like Link-11, or a high rate digital voice link, (16 Kbps).

The multi-function CNI requirements of the host platform are generally met with a number of FCs operating within one MC. There are up to 512 BCs available on each MC, for use in the composition of FCs. These levels are depicted pictorially in Figures 49, 50, 51.

The capacity of a MC is 300 Kbps. Specific terminal channel capacities are tailored to the particular operational needs. The channels are designed to meet C3 requirements and they are tailored efficiently in terms of access rate (message start rate), data rate (number of BCs) and composite channel

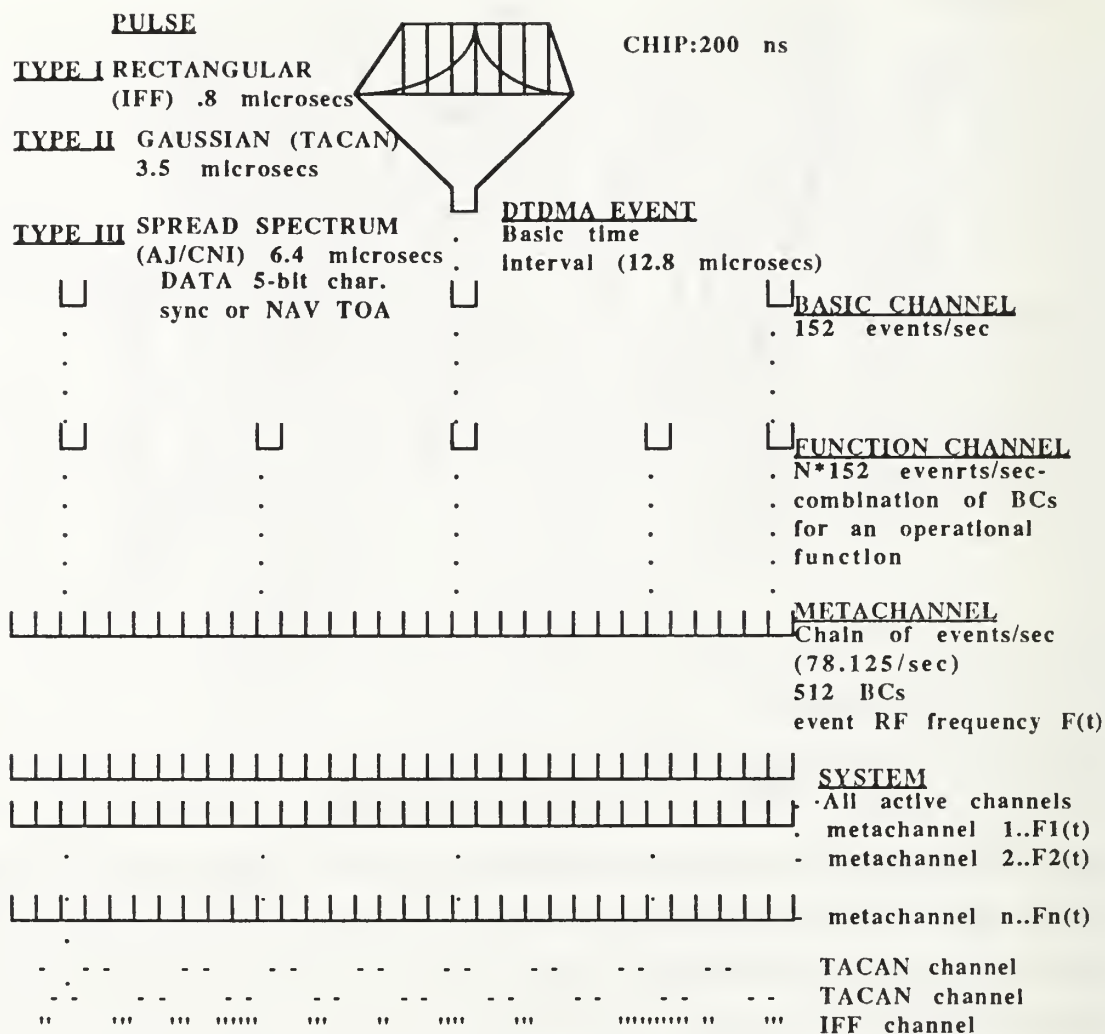


Figure 48. DTDMA: channel architecture [Ref. 15: p.6].

grouping (composition of multi-function nets), for specific operationally driven communication needs.

B. DESCRIPTION OF THE SYSTEM DESIGN

The system is essentially designed to operate at four levels. These levels identify respective terminal classes. While JTIDS has basically four classes of TDMA (DTDMA) terminals, the proposed system is slightly different. The respective modifications are shown in Table 5 and their specifications are given in Table 6.

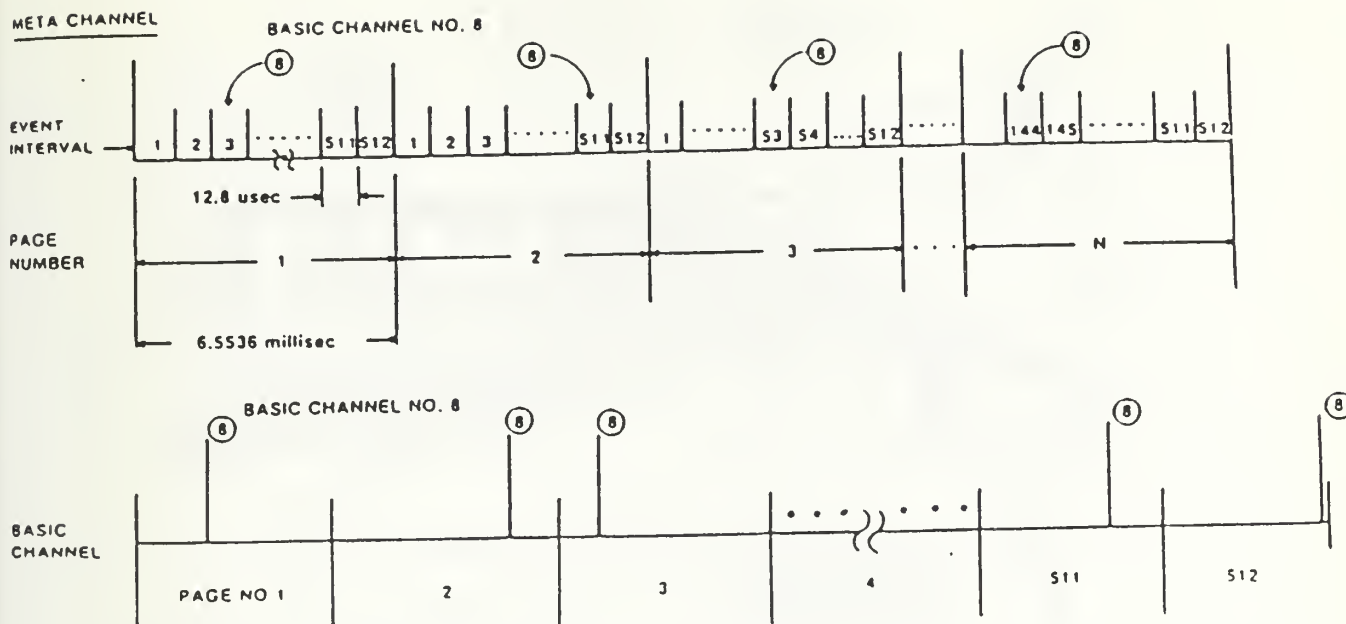


Figure 49. DTDMA: Basic Channel to Metachannel

Table 5. PROPOSED SYSTEM TERMINAL MODIFICATIONS (VS) JTIDS TERMINALS

COMMAND LEVEL	JTIDS	PROPOSED SYSTEM
Surface/airborne C3 centers	Class 1-1A (surface centers/ airborne E-2C's)	Class 1 Ground C3 centers
Subsidiary commands	-	Class 2AM-MPA's
Individual units	Class 2-2A ships(frigates)/ fighter aircraft	Class 2-2AF ships(frigates)/ fighter aircraft
Manpack-vehicular	Class 3 ground forces	Class 3 ground forces

A JTIDS based design is drawn from a baseline terminal architecture. This particular baseline terminal consists of three major parts.

- receiver/transmitter for special processing and conversion.
- Digital Data Processor (DDP) for real-time control and message processing.

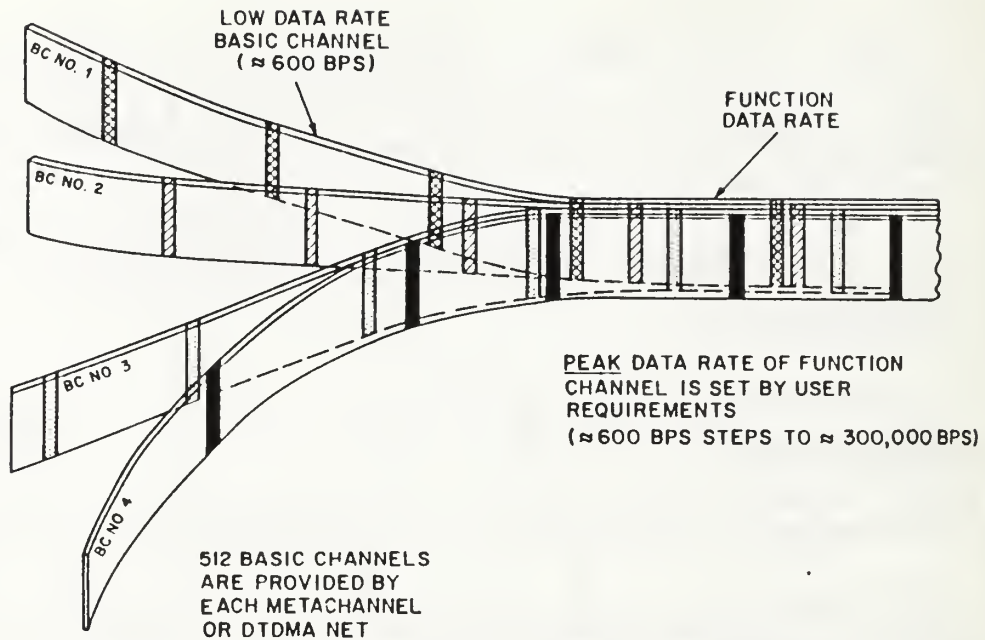


Figure 50. DTDMA: Composition of Function Channel [Ref. 17: p.37-8].

- interface unit to tailor the terminal to the platform being used.

A functional configuration of this design is shown in Figure 52. This design is essentially the Class 2 baseline for the Tactical Aircraft (FAC) and it also forms the basis for all the other versions. A different representation of this baseline design together with both the extended airborne version for the E2-C and the AWACS aircraft, and the basic shipborne version designed to be installed on several ships, are shown in Figures 53, 54, 55. The proposed system is based on the JTIDS class 2 baseline terminal which becomes the system's basic airborne version for the FAC (Fighter Aircraft). A slightly modified airborne version close to the baseline is intended for the MPA. The basic shipborne version remains the same, but the large ground command version has such features as the class 1 terminal originally designed for the airborne C3 centers, such as those in E-2Cs.

The proposed modifications will add little engineering effort and cost since the JTIDS-baseline terminals are modularized. The basic differences are in the size and weight considerations needed for the MPA version and the use of a

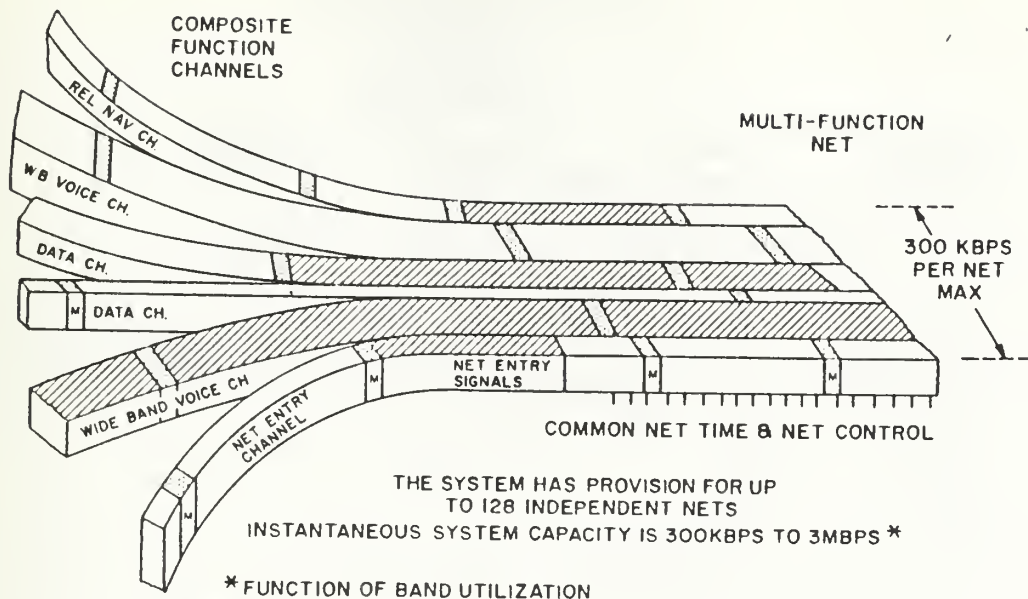


Figure 51. DTDMA: Composition of a Multi-function net [Ref. 17: p.37-9].

Table 6. TERMINALS FUNCTIONAL CHARACTERISTICS

CHARACTERISTIC	CLASS 2	CLASS 1A	CLASS 1
Data rate, Kbps (Tx/Rx/TX-Rx)	80/140/160	160/140/160	280/280/300
Open message start rates, Hz (sync/sync-data)	50/100	50/100	100/200
Number function channels	64	64	128
Number voice channels	2	4	10
Number conferencing voice channels	1x3	1x3	2x3
TACAN	Yes	Yes	N/A
Number metachannels	6	8	16
Quaternary function channel	2	2	4
Number of users	512	512	512
Antennas	2	1	2
RF power, Watts (normal/military/burnthrough)	200	200/400/ 1600	200/400/ 3200
Range, NM	300	300	300

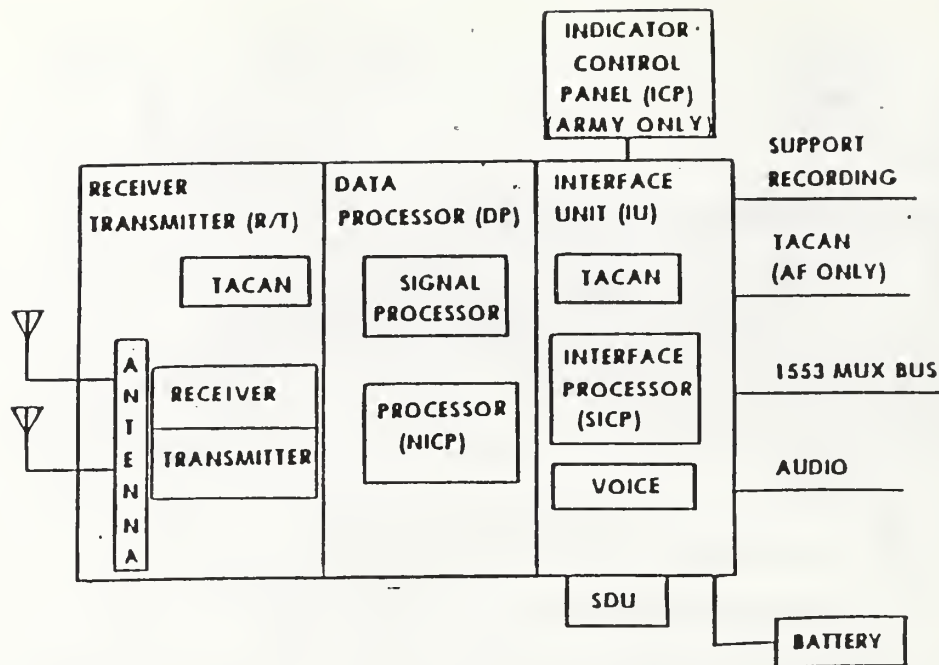


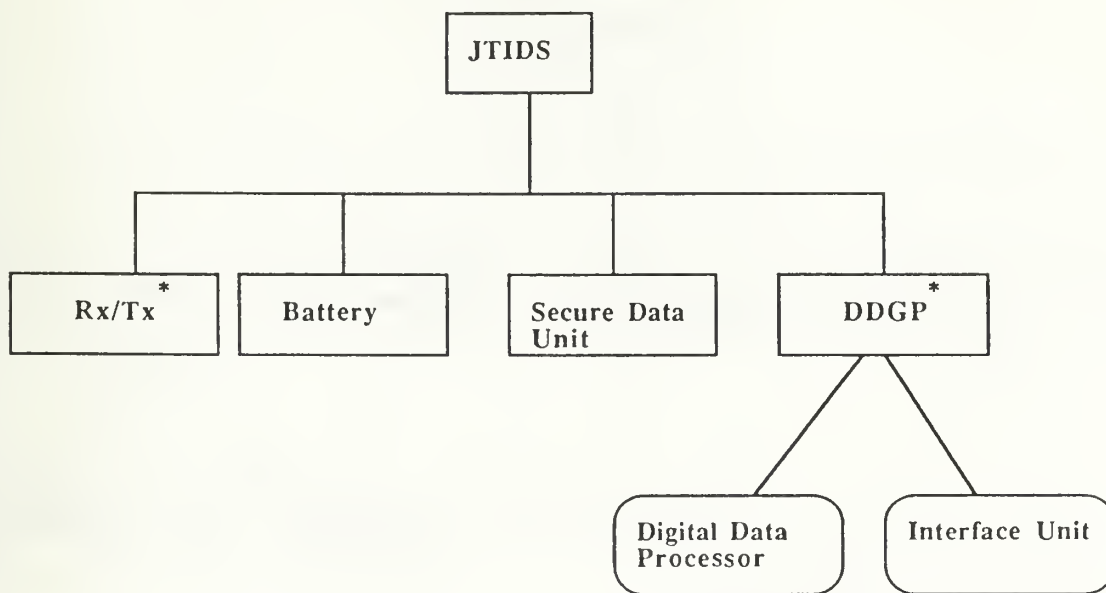
Figure 52. Baseline Terminal: functional configuration [Ref. 18: p.7].

different high power amplifier (HPA) than the one used in the version designed for the E-2Cs. The cost effectiveness of this modification is further examined in the last chapter.

1. Receiver/Transmitter (Rx/Tx)

A block diagram of the baseline class 2 terminal with the receiver/transmitter and the digital processor modules is shown in Figure 56. It provides the signal processing necessary for the realization of the JTIDS and TACAN functions. It is basically a multi- function, dual conversion heterodyne receiver, capable of simultaneous operation on four independent channels. It can process the spread spectrum continuous phase shift modulated (CPSM) JTIDS pulses, the conventional TACAN gaussian and the rectangular IFF pulses too. The JTIDS pulses are provided by the CPSM modulator and up converter stages. Frequencies are supplied by four frequency synthesizers which are capable of providing frequencies in increments of 1 MHz across the whole range of the operating band (255 MHz). Down conversion is done by a

NAVY CLASS II BASELINE
(FIGHTER AIRCRAFT)



*DDGP: Digital Data Group Processor

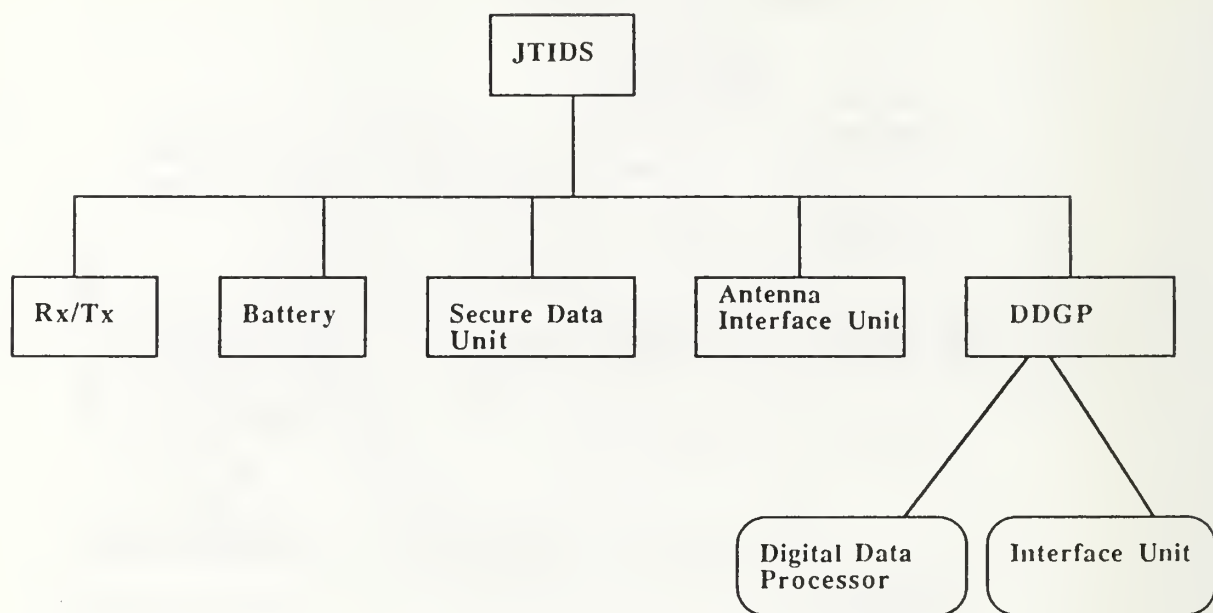
*Rx/Tx: Receiver/Transmitter

Figure 53. Navy Class II Baseline: airborne (FAC) version.

quadrature L-band converter which shifts the incoming L-band signal down to the IF frequency at 70 MHz after being fed with the 280 MHz output of a frequency source module. Channelization is achieved by using four IF amplifiers with approximately 80 dB gain. Matched filter detection is used for the correlation of the spread spectrum pulses.

The TACAN receiver/synthesizer, power amplifier and processor are independent from the JTIDS functions. One of the channels is exclusively used for reception of the TACAN pulses. The same is also done for IFF reception. The Rx/Tx varies from terminal class to terminal class. The functions of the Rx/Tx are summarized in Table 7 [Ref. 19: p.38-3].

NAVY CLASS IIN (E-2C)



*DDGP:Digital Data Group Processor

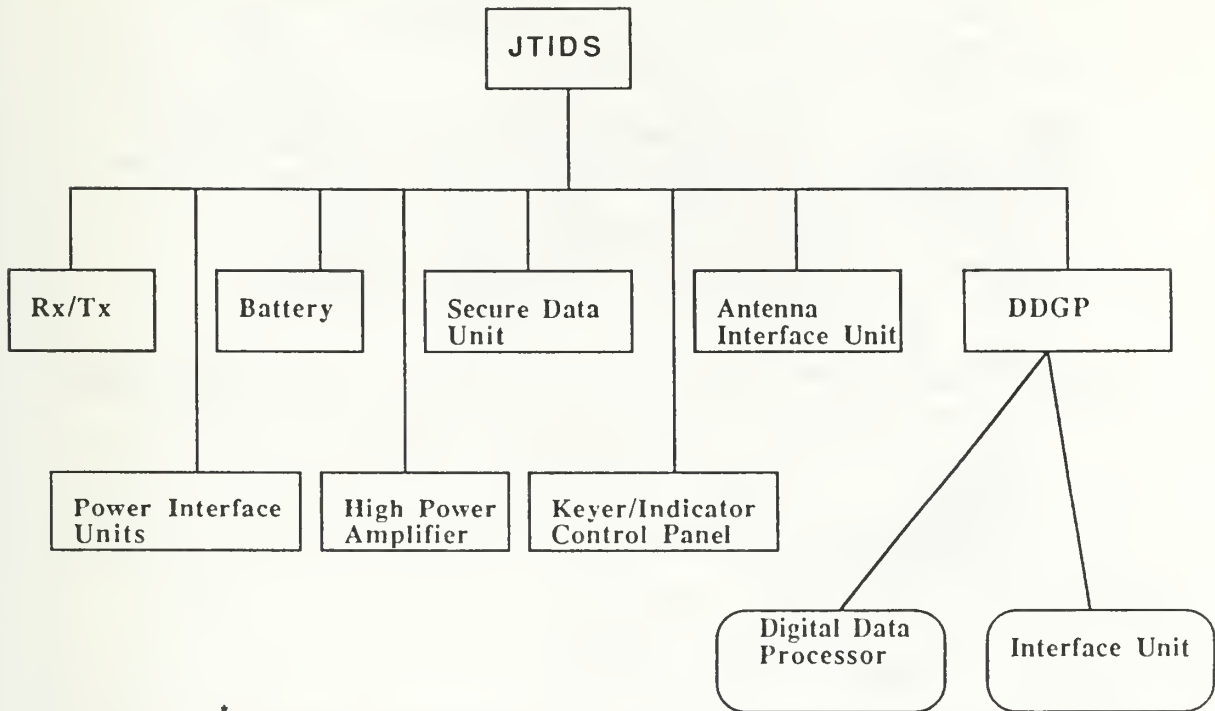
*Rx/Tx: Receiver/Transmitter

Figure 54. Navy Class IIN terminal: airborne C3 version (E-2C, E-3A).

a. The Data Processor

This module performs real-time control and signal processing for the JTIDS message transmission and reception. The whole process is depicted pictorially in Figure 57 [Ref. 19: p.38-4]. The transmitted data enter the buffer area at point (1), and then are sent to a Reed-Solomon encoder module for A-J protection (enhancement), at point (3). The coded data then are sent to the modulator stage at point (4) where each five bit character is impressed on a 32-chip digital word. Then the data is further processed through the pseudorandom source at point (5), where two basic functions take place. First, the data is mixed with an equal number of PR bits by modulo-2 addition (PR spreading) and is sent to the RF modulator in the transmitter at point (6),

NAVY CLASS (SHIP)



* DDGP: Digital Data Group Processor

*Rx/Tx: Receiver/Transmitter

Figure 55. Navy Class II Terminal: shipborne version.

where it phase modulates the RF carrier at the transmission time. Second, the PR source provides a PR bit stream to the frequency-time hopping control at point (7), which then provides a pseudorandom carrier frequency and time-hopped transmission schedule at point (8). The process results to the generation of RF pulses at L-band, at point (9). These pulses have the following high AJ properties:

- data collection/buffering.
- error protected.
- PR spread.
- time-frequency hopped.

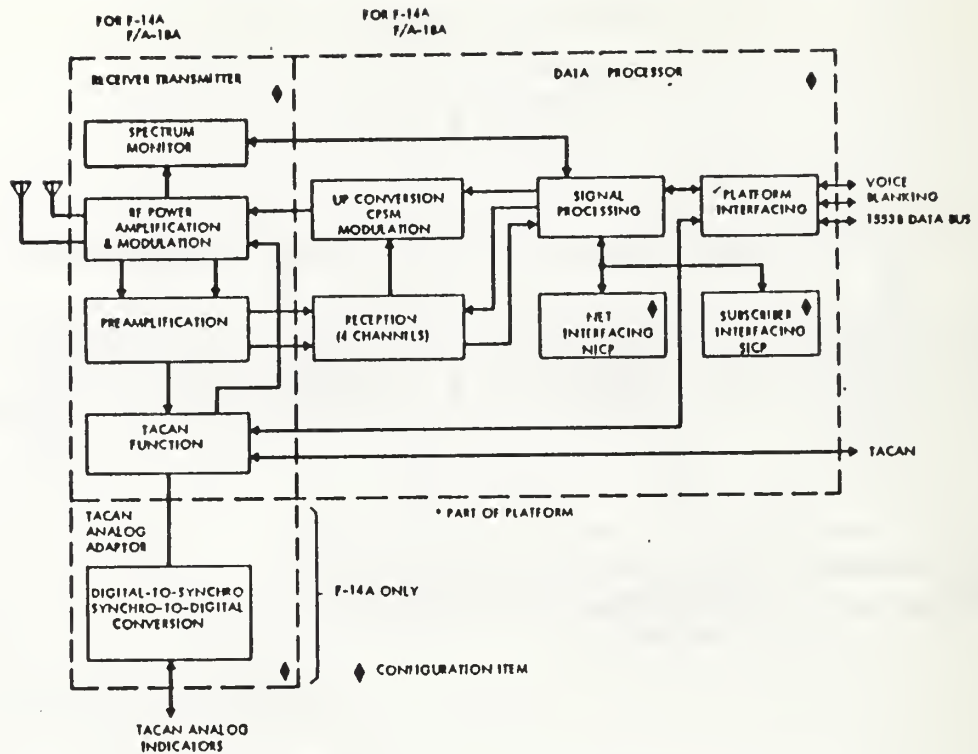


Figure 56. Class 2 Baseline Terminal: functional block diagram [Ref. 18: p.7].

Table 7. CLASS 2 TX/RX FUNCTIONS

Low noise preamplification
JTIDS and TACAN transmit
TACAN transmit gaussian shaping
Signal amplification (JTIDS: 200W/ TACAN:500W)
TACAN input control
TACAN processing
TACAN antenna selection
Spectrum monitoring
Rx/Tx bit status reporting
Transmit and receive filtering

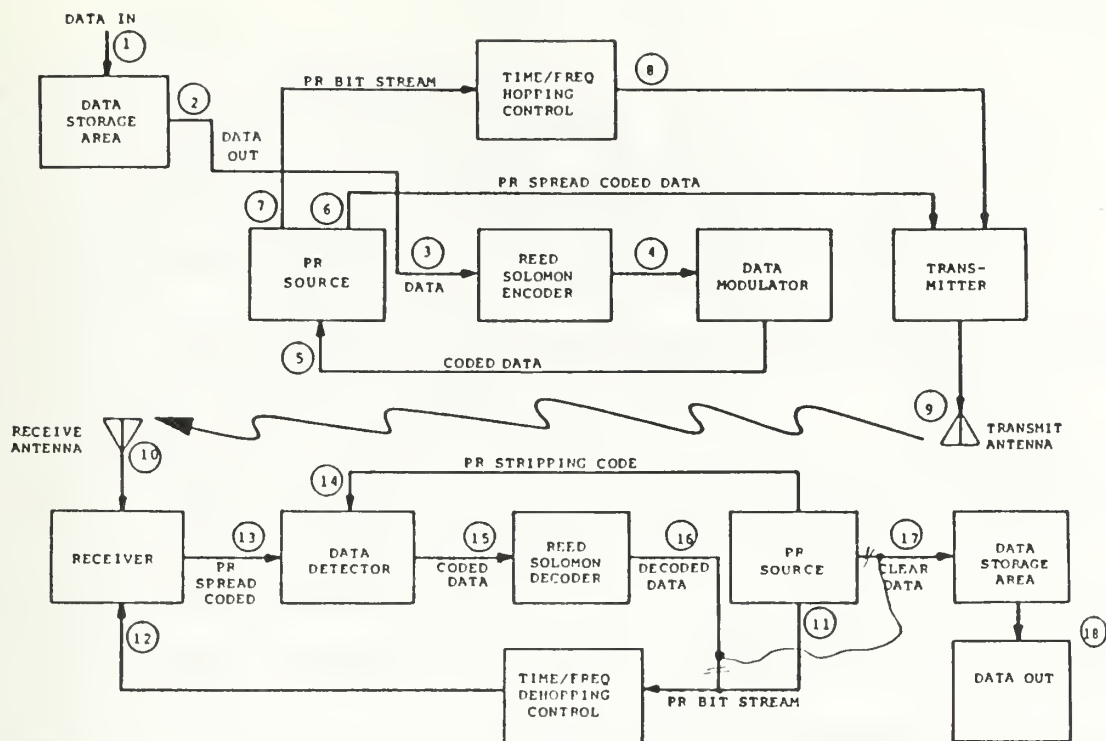


Figure 57. Digital Data Processor: Tx/Rx signal flow [Ref. 19: p.38-4].

The transmitted pulses are then remotely received at point (10). An internally generated PR bit sequence at point (11) serves as a reference to the time/frequency de-hopping control stage for the de-hopping of the received pulses at point (12). The de-hopped output is then fed into the data detector which removes the PR code at point (14) and demodulates the 32 chip data word. The demodulated 5-bit data character is decoded at point (15), and the clear data are fed into the buffer area for temporary storage and final routing to the intended interface (points (17), (18)) [Ref. 19: p.38-3, 38-4].

b. Time scheduling/PR control

A modularized diagram of the digital processor is shown in Figure 58.

Time scheduling is realized by the Schedule Processor, the Time Ordered Lists, the Terminal Clock, and the Composite Storage Modules. The schedule processor basically orders the transmission and the reception

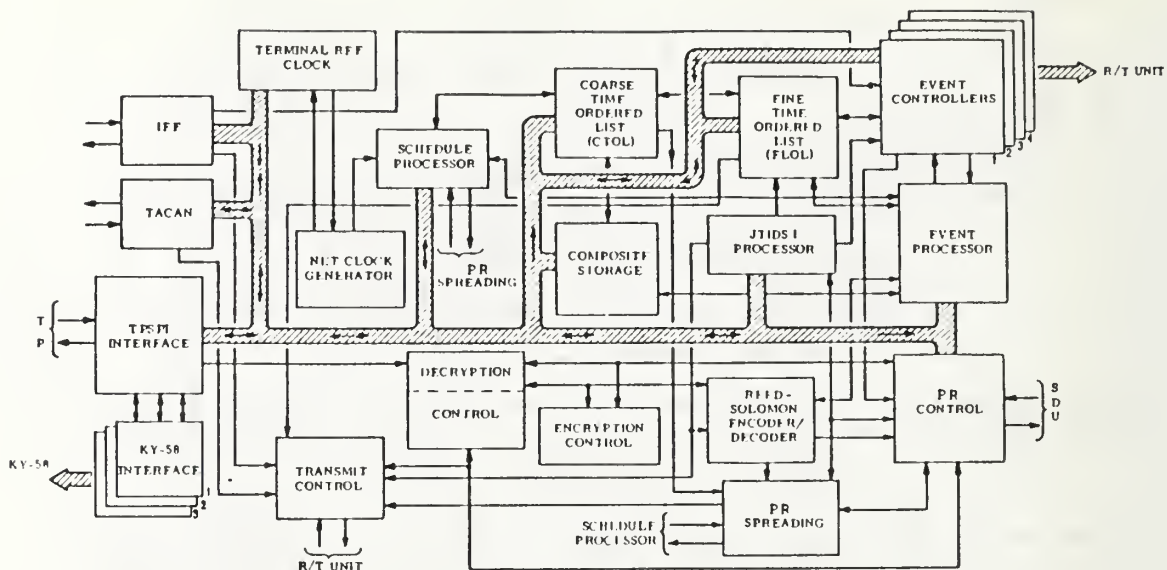


Figure 58. Signal Processor: modularized diagram [Ref. 19: p.38-4].

events. These time ordered events are loaded into a Coarse time ordered list which is a mechanism that maintains proper event order. The events are stored there properly formatted until they are passed to the Fine time ordered list for priority routing and final assignment to one of the four receiving channels or to the transmitter [Ref. 19: p.38-4,38-5].

PR control is achieved by suitable circuitry that includes functions such as PR spreading, encryption/decryption and PR control stages. It provides the necessary PN sequences for the transmitted pulse modulation and the reference PN sequence for the despreading of both the received pulses and the required carrier frequencies.

The Transmitter/Receiver (Tx/Rx) control is realized by the Event Controller, Event Processor and the Reed Solomon Encoder/Decoder Modules. On transmission, the messages from the data memories are passed by the Transmit Controller to the R/T unit, suitably screened for conflict avoidance. Similarly, on reception, the Event Controller takes the Time Ordered List and assigns one of the four available receivers. The received Time of Arrival

Table 8. CLASS 2 DIGITAL PROCESSOR FUNCTIONS

Channel control
Transmit CPSM shaping
Transmit/Receive up conversion
JTIDS antenna selection
Preamble detection
Sync processing
TDMA message processing
Chip tracking
Reed/Solomon encoding-decoding
Header and RTT processing
CCSK encoding
Data routing
Message buffering
Conflict resolution
Signal detection/TOA processing

(TOA) is fed back to the Event Controller by the R/T unit. Data and event information are fed to the processor which then adjusts for any tracking error. Non-encoded messages are directly passed to the Decryption Control Circuitry, where encoded messages are passed through the Reed-Solomon encoder to the Decryption control. TACAN and IFF functions are provided as well [Ref. 19: p.38-5]. The main functions of the DP are summarized in Table 8 [Ref. 12: p.4].

2. The Interface Unit (IU)

The IU basically provides a standard interface between the host platform and the DP. It performs the necessary transfer of data between the platform and the terminal. It provides channel configuration and mode control, two way data flow control and data buffering with external interfaces such as LINK 4A, Link 11, TACAN and Relative Navigation systems. It also provides interface

with the Control and the Display unit. It contains the Network Interface Digital Computer (NIDC) and the Subscriber Interface Digital Computer (SIDC). The functions performed by these computers are shown in Figures 59, 60.

A summary of the tactical functions provided by a typical class-2 terminal is given in Table 9 [Ref. 19: p.39-2].

Table 9. JTIDS II TACTICAL TERMINAL FUNCTIONS

DTDMA FUNCTIONS
(1) 16 Kbps direct voice channel
(1) 16 Kbps direct/relay voice channel
Link 4A digital data link
Link 11 digital link relay
Relative navigation
Loading rate (Rx:50 Kbps, Tx:40 Kbps)
Capacity (Rx:70 Kbps, Tx:50 Kbps)
TDMA FUNCTIONS
Relative navigation interoperability
Net management processing
Round trip time measurement
Capacity (Rx:70 Kbps, Tx:70 Kbps)
Airborne TACAN (Tx/RX, Rx and air-to-air modes)
IFF transponder modes 1, 2, 3/A, C and 4)

C. SYSTEM LINK ANALYSIS

In this section, an effort is made to examine a JTIDS-based design from the perspective of a communications link between a transmitting and a receiving terminal. A typical generalized model for this link is shown in Figure 61.

The first step is the calculation of the system baseband bandwidth (B), which is essentially the information bit rate (R_b), irrespective of the encoding, spreading and modulating of the baseband signal. As it has already been explained in previous sections, the basic JTIDS signal consists of three 75-bit

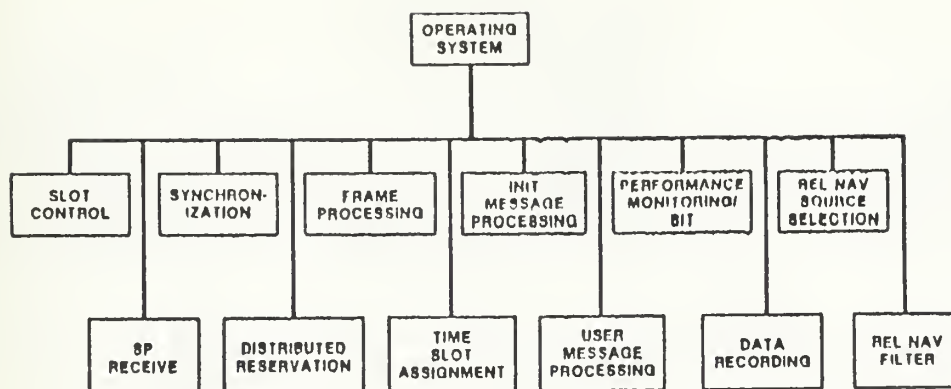


Figure 59. NIPC: functional block diagram [Ref. 18: p.10].

words in every time slot, out of a total of 128 time slots available in every second. So,

$$R_b = (\text{No of words}) * (\text{No of bits/word}) * (\text{No of time slots/sec}) = 28,800 \text{ bits/sec}$$

Then, based on the system specifications it is known that the RF bandwidth is 3 MHz. So, the system processing gain due to the PN spreading can be subsequently calculated below, as follows:

$$\text{Processing Gain: } G_{\text{pdB}} = 10 \times \log(3\text{MHz}/28.8\text{KHz}) = 20.18\text{dB}$$

This parameter is an important feature of the system, and is used for the jamming margin calculation in the next section.

The bandwidth available for hopping is 237 MHz, as the frequency ranges from 969 MHz to 1206 MHz. By excluding a total of 84 MHz reserved for IFF and guard bands, the total usable bandwidth for hopping is 153 MHz. Then the available number of pseudorandomly hopped channels of 3 MHz each, is 51. These 51 channels correspond to different carrier frequencies available for the pseudorandom transmission of the signal. The hopping pattern is also shown in Figure 61.

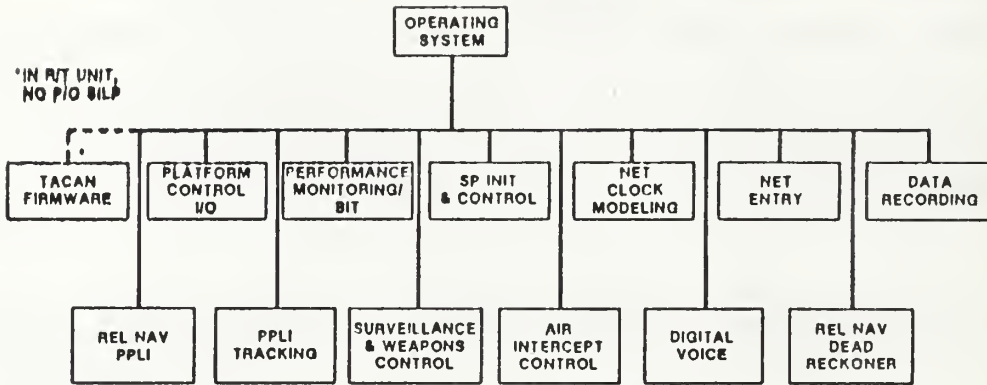


Figure 60. SIPC: functional block diagram [Ref. 18: p.10].

In attempting to set up a model for the system link analysis, the typical equation used is:

$$\frac{G_t \times G_r}{L_p} = \frac{f_c^2 \times A_{et} \times A_{er}}{D^2 \times c^2}$$

where,

f_c : the carrier frequency used for transmission

G_t, G_r : antenna gains for the Tx/Rx

D : the distance between the Tx and the Rx at which potential reception is possible

c : the speed of light

L_p : propagation loss along the transmission path including free space loss

A_{et}, A_{er} : effective antenna areas for the Tx/Rx

Assuming the following typical parameters for an air-to-ground link, and varying the carrier frequency over the entire hopping range, the net effect on the communication link performance can be calculated as below.

Let:

- maximum system range $D = 300$ nautical miles = 555.6 Km.

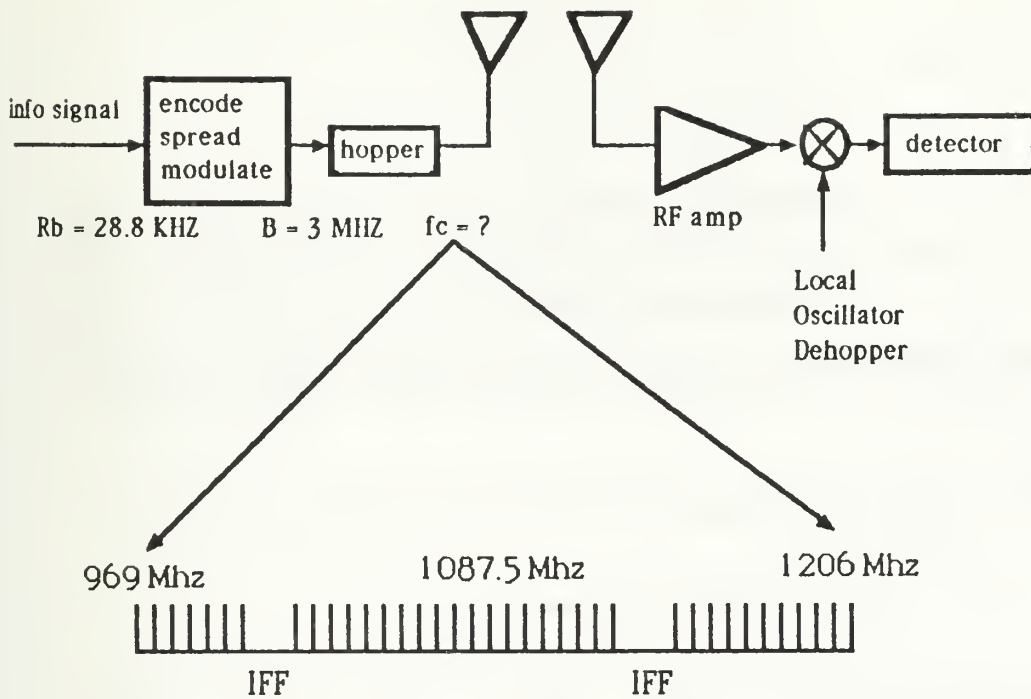


Figure 61. Link Model: Tx/Rx link configuration.

- air antenna gain G_r (Tactical Fighter) = 0 dB.
- ground antenna gain G_t = 11 dB.

(a). For $f_{cLower-Limit} = 969 \text{ MHz}$

$$\frac{G_t \times G_r}{L_p} = -153.81 \text{ dB}$$

(b). For $f_{cMean} = 1087.5 \text{ MHz}$

$$\frac{G_t \times G_r}{L_p} = -152.81 \text{ dB}$$

(c). For $f_{cUpper-Limit} = 1206 \text{ MHz}$

$$\frac{G_t \times G_r}{L_p} = -151.81 \text{ dB}$$

The net result is approximately 1 dB improvement in communication performance as the carrier frequency is increased step by step to the upper possible limit. This effect can be evaluated also in terms of the resultant maximum system communication ranges as a function of the carrier frequency, based on the following formula:

$$D = \sqrt{\frac{P_t \times G_t \times G_r \times c^2}{16 \times \pi^2 \times K \times T \times B \times L_a \times \text{SNR} \times f_c^2}}$$

which is another version of the link equation, where:

K: Boltzman's constant = 1.38×10^{-23} J/Kelvin

B: bandwidth available for hopping or bandwidth of the RF amplifier at the receiver

T: effective system noise temperature in degrees Kelvin

SNR: Signal-to-Noise ratio at the input of the demodulator

L_a :attenuation loss

Assuming a tactical air-to-surface link between a tactical fighter and a ship, and certain typical parameters that are applicable in this case, the net effect of the carrier frequency variation on the communication range can be calculated as follows:

- maximum SNR threshold for a digital link = 15 dB.
- $T_a = 290$ K.
- $T_e = 900$ K.
- maximum transmitter power $P_t = 200$ Watts.
- $G_t = 7$ dB.
- $G_r = 0$ dB.

- ignoring attenuation loss, $L_a = 0$ dB.

Then, the maximum communication range D becomes:

(a). For $F_c = 969$ MHz $D = 624$ Km = 336 nm

(b). For $F_c = 1087.5$ MHz $D = 555.6$ Km = 300 nm

(c). For $F_c = 1206$ MHz $D = 501$ Km = 270.5 nm

The conclusions drawn from these calculations are:

- There is about 11 percent reduction in the system range as the frequency is increased from 969 to 1206 MHz (the range is inversely proportional to the frequency).
- At 969 MHz the maximum achievable communication range of 336 n. miles appears to be greater than the system's maximum range of 300 n. miles (beyond line-of-sight).
- no atmospheric attenuation loss was considered here.
- the use of the upper frequency limit (worst case carrier), reduces the average achievable range by about 11 percent. Since reception of the entire signal across all the 51 hopping channels is necessary, this can be considered the most suitable (worst-case) carrier frequency for system configuration and deployment planning.

D. DEPLOYMENT CONSIDERATIONS

The system design is such that it provides a high level of connectivity among the decision making commanders, their scattered sources of surveillance and their weapon systems. Since the system operates in a broadcast mode, its connectivity implies the use of omnidirectional antennas. Also the fact that the system operates in the UHF band limits its communication range to line-of-sight. Therefore, coverage beyond line of sight is practical using relays. This is possible because all terminals can be configured to act in a relaying mode. This implies a penalty for the relay unit as its terminal capability is reduced by half because of the necessity of receiving in one time slot and transmitting during another.

A typical naval tactical deployment in a closed-sea area of operations may include a Joint Task Force with the following composition;

- several Guided Missile Destroyers/Frigates with helicopters.
- several attack submarines.

- several landing and supply ships.
- sufficient air reconnaissance and air-striking support.
- a number of MPAs in a surveillance role.

Obviously the Task Force commander needs total connectivity. To maintain this level of connectivity, a significant number of tactical voice, data and relay circuits is required. Moreover, this task force may be spread in transit over an area as large as 300 n. miles in radius, presenting communication problems even to the most advanced systems. Many of the tactical circuits are UHF and this limits them to line of sight distances (15-20 n.m between surface platforms and 200-300 n.m surface-to-air for approximately 30,000 ft altitude). This is a severe limitation since dedicated relay units are needed in order to maintain a satisfactory level of information exchange.

A JTIDS-based design appears to provide extremely high capacity for this kind of multi-net situation. If all units of the task force are equipped with JTIDS terminals, then by combining similar functions the big number of tactical circuits mentioned before will be reduced significantly. Net management then will be more efficient and all the task-force units can act as relays. Therefore wide area coverage will be achievable with jam-resistant real-time data transfer in any tactical situation (LR AAW or SR ASW).

But system line-of-sight range limitations imply the use of an HF net (Link 11-Link 14) or the existence of a dedicated airborne relay for wide area coverage. Although Link 11 is an HF data link ensuring beyond line of sight information exchange, it is very vulnerable to enemy's jamming and of limited data capacity. And the second option of an airborne relay appears not to be very attractive.

Given the capabilities and the limitations of the system, the task force commander has to configure his system in the most efficient way in order to carry out his mission. There are so many different tactical situations that a unique configuration is not possible. For instance, a surface unit (ship) will communicate with a Fighter Aircraft through an air communications path, normally under good weather conditions. On the other hand, a good command

station will communicate with a surface unit under all weather conditions, over rough terrain and sea-state conditions.

The efficient configuration is a function of the power settings available and the type of the antennas used. The commander can operate on these parameters and configure the system in the best way. But there is a third parameter, attenuation loss along the communication path, which is not adjustable. More specifically, the attenuation loss per kilometer is more appropriate for consideration since path losses, expressed in decibels, are directly proportional to the distance in the case of uniform attenuation. Several factors contribute to this loss, such as weather conditions and terrain (which is mainly a matter of line of sight and antenna-height placement which affects the degree to which the communications range is enhanced). Based on these factors, tactical plans and equipment configurations will depend on the expected system performance.

A system communication range analysis for a variety of different configurations and several typical values of L_a/km encountered in UHF propagation provides useful information to the commander in determining its most efficient configuration. Such a generic analysis, for three different types of antennas, three different power settings and a wide spectrum of possible L_a/km values, is performed below.

The particular scenario employs an air (aircraft)-to-surface (ship) communications link, an air (MPA)-to-ground (command center) communications link and an air (MPA)-to-surface (ship) surveillance link. Typical antenna gains used are:

- $G_t, G_r = 11\text{dB}$, for a Ground Command Center.
- $G_t, G_r = 7\text{dB}$, for a ship.
- $G_t, G_r = 0\text{dB}$, for an MPA, FAC.

Typical power settings:

- $P_t = 1600\text{W}$, for Ground Command Center.
- $P_t = 400\text{W}$, for an MPA.
- $P_t = 200\text{W}$, for a FAC and a ship.

Also, the attenuation La/km varies depending on the following:

- Terrain.
- Vegetation.
- Sea-state.
- Weather.
- Antenna height and placement.

The methodology used here is based on the formula that gives the link range:

$$D = \frac{c}{4 \times \pi \times Fc} \times \sqrt{\frac{Pt \times Gt \times Gr}{k \times T \times B \times La \times SNR}}$$

Using the expression of La as a function of distance D for a one way link with uniform attenuation, we have: [Ref. 20: p.459].

$$La = e^{\alpha \times D} \text{ or } (La)_{dB} = \beta \times D$$

where,

$$\beta = 4.34\alpha$$

β : is the specific attenuation in dB/km

D: is range in km

If the uniform attenuation only applied to a fraction of the range D, then that fraction would be used in the exponent of the equation for La. If the attenuation were non-uniform, then La would be calculated by integration. For simplicity, in the following, only the case of uniform attenuation over the entire range, D, will be considered.

Substituting back to the previous expression we get:

$$D = \frac{c}{4 \times \pi \times Fc} \times \sqrt{\frac{Pt \times Gt \times Gr}{k \times T \times B \times e^{\frac{\beta \times D}{4.34}} \times SNR}}$$

Then, by substituting the parameters assumed before, we get the final expression for the link range as a function of the attenuation:

$$D = \frac{C}{e^{\frac{\beta \times D}{8.68}}}$$

$$\text{where, } C = \frac{c}{4 \times \pi \times Fc} \times \frac{(Pt \times Gt \times Gr)^5}{(k \times T \times B \times SNR)^5}$$

And for this substitution there are three cases:

(a). *Air-to-surface link (FAC-SHIP)*

$$D = 500 \times e^{\frac{-\beta \times D}{8.68}}$$

(b). *Air-to-ground link (FAC-COMMAND CENTER)*

$$D = 796 \times e^{\frac{-\beta \times D}{8.68}}$$

(c). *Air-to-ground link (MPA-COMMAND CENTER)*

$$D = 1124.64 \times e^{\frac{-\beta \times D}{8.68}}$$

(d). *Air-to-surface link (MPA-SHIP)*

$$D = 709.4 \times e^{\frac{-\beta \times D}{8.68}}$$

where,

D is measured in km

All the four curves describing the above cases follow the same declining pattern. The first case is typically shown in Figure 62. Obviously different power settings and antenna gains will yield different ranges, but in any case the same generic curve pattern will result. The exponential reduction in achievable range becomes apparent as the attenuation is increased from a line-of-sight, free space, clear-sky path model (0 db/km) to a near-surface path over rough terrain (0.01 dB/km).

This generic system range analysis applies to any kind of geographic area of operations, provided that the attenuation loss factor can be precisely measured. This could then be used by the commander in forming his tactics, such as units' deployment and spacing. Consequently, this will aid him in establishing reliable communications, without relying too much on the relay capability of the system or on HF links.

E. SYSTEM JAMMING CONSIDERATIONS

Considering potential ways of jamming a JTIDS-based system, there are two possibilities. Barrage or smart jamming. These will be further analyzed in the following sections.

1. Barrage Jamming

By barrage jamming, the jammer (spaceborne or airborne) is forced to transmit considerable power in order to cover the huge spectrum of a JTIDS signal, and to overcome the system's jamming margin. This signal is a hybrid combination of a pseudorandom (PN), frequency hopping (FH) and time hopping (TH) pulse and so its jamming margin is enhanced by the respective

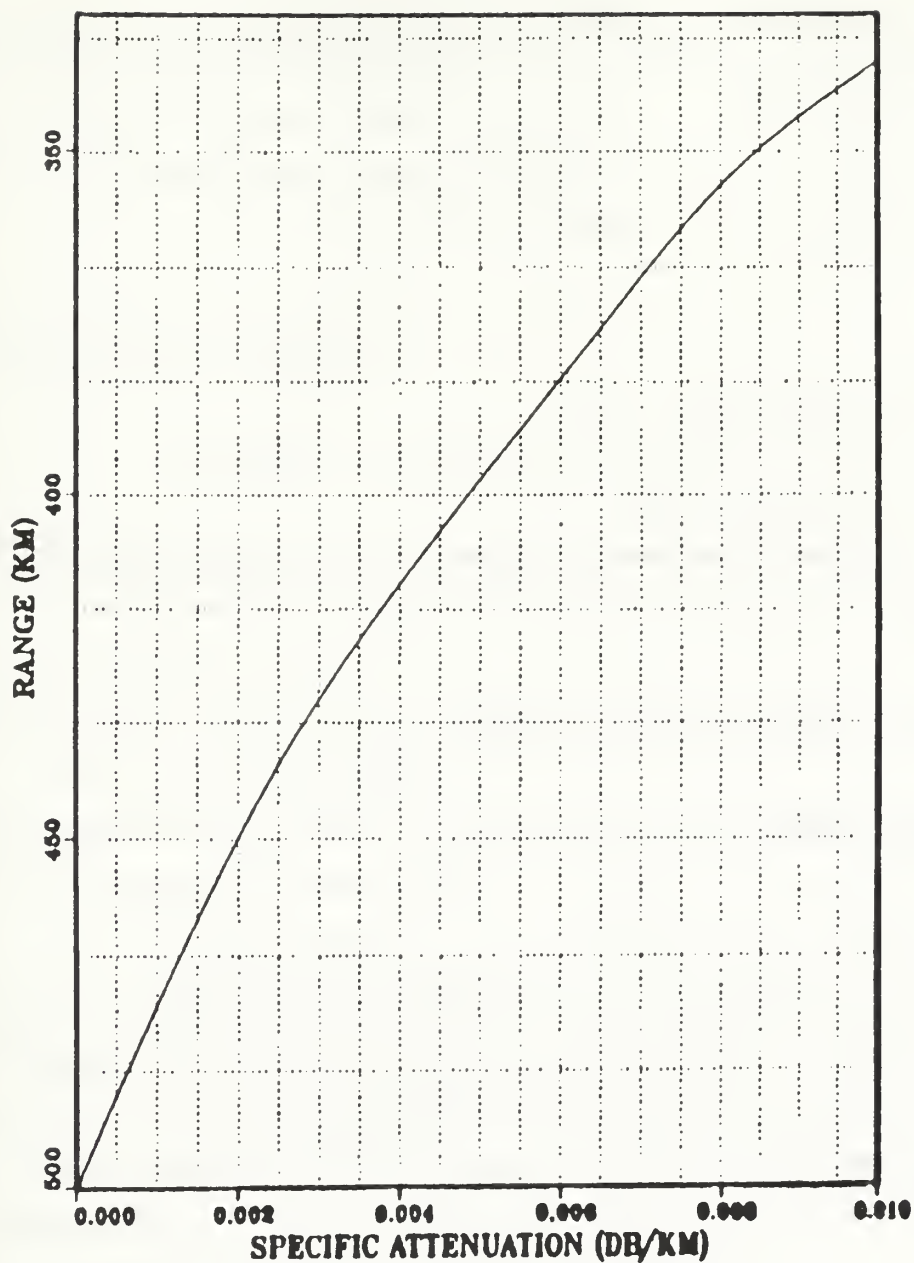


Figure 62. Air-to-Surface Tactical Link: range vs specific attenuation.

processing gains of each one of these particular techniques. The combined system processing gain is therefore the sum of these individual gains. This margin is expressed by the following equation for a given processing gain G_p ,

a desired S/N at the output of the receiver demodulator and a particular system loss factor L_{sys} .

$$Mj = J/S = Gp_{hybrid} - S/N - L_{sys}$$

where,

$$Gp_{hybrid} = Gp_{PN} \times Gp_{FH} \times Gp_{TH}$$

$S/N = 15\text{dB}$ maximum desired SNR for a digital link

$L_{sys} = \text{typically } 1 \text{ dB}$

So, based on the theory of chapters 2 and 3 the processing gain of the system is:

$$Gp_{hybrid}(\text{dB}) = 20.18 + 17 + 51.9 = 89\text{dB}$$

Which gives:

$$Mj = J/S = 89 - 15 - 1 = 73\text{dB}$$

This is an extremely large figure for the jammer to overcome before any effect can be observed on the system's receiver performance, when smart jamming is intended. When barrage jamming is chosen, the effective system Gp is decreased to the figure gained by the spreading and the hopping factor. This Gp is about 38 dB with a subsequent jamming margin for the system of about 22 dB.

A realistic scenario is two JTIDS-equipped ships trying to communicate with each other over a JTIDS link at practical distances up to 20 nautical miles (NM) (line-of-sight), under optimum conditions of no sea-surface attenuation. Then an airborne jammer, rather than a shipborne one, is regarded as the most likely threat against the system receiver due to its relative unpredictability and higher jamming distance capability. Also, if this jammer attempts to operate under optimum conditions (clear skies and no atmospheric losses)

then based on the assumed parameters given below, the jamming equation can be set for this case (see Figure 63).

- shipborne antenna gains = 7dB.
- airborne directional jammer antenna = 20dB.

$$J/S = \frac{(P_j \times B_r) \times G_{jr} \times G_{rj} \times L_{pjr} \times L_{ajr}}{P_t \times G_{tr} \times G_{rt} \times L_{ptr} \times L_{atr}}$$

where,

$P_j \times B_r$: total jamming power P_T needed to cover the receiver's bandwidth (P_j : power/Hz, B_r : receiver's bandwidth)

P_t : transmitter power

G_{jr} , G_{rj} : antenna gains along the jammer-to receiver and the receiver-to-jammer paths

G_{tr} , G_{rt} : antenna gains along the Tx-Rx and the Rx-Tx paths save

L_{pjr} , L_{ptr} : free space losses for the jammer-receiver

and the Tx-Rx paths

L_{ajr} , L_{atr} : propagation loss for the

jammer-receiver and the Tx-Rx paths
(here ignored)

Then, the previous jamming equation can be rewritten:

$$J/S(dB) = (P_j \times B_r)(dB) + G_{jr}(dB) + G_{rj}(dB) + L_{pjr}(dB) - (P_t)(dB) \\ - (G_{tr})(dB) - (G_{rt})(dB) - L_{ptr}(dB)$$

where, The free space loss can be expressed as :

$$L_p(dB) = \left(\frac{4 \times \pi \times D}{\lambda} \right)^2$$

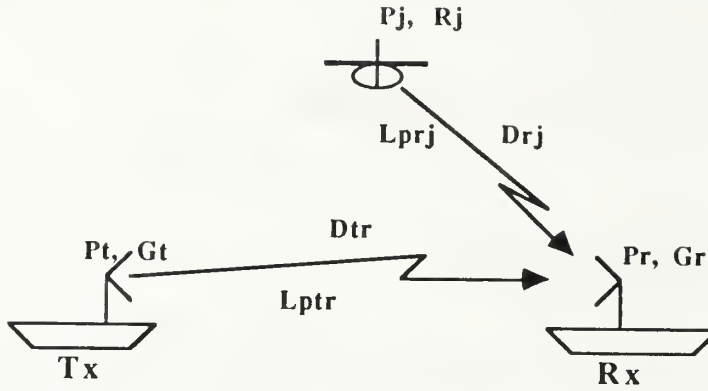


Figure 63. Air-to-Surface Jamming: depiction of a typical scenario.

where,

D: distance between the jammer and the receiver or the
Tx-Rx

λ : operating wavelength

The total jamming power required to overcome the receiver jamming margin, when barrage jamming is intended, can be expressed as a function of the jammer-to-receiver distance.

$$J/S(dB) = P_T(dB) + 27 + 20 \times \log \frac{c}{f} - 20 \times \log(4 \times \pi) - \\ - 20 \times \log Drj - 23 - 14 - 20 \times \log \frac{c}{f} + 20 \times \log(4 \times \pi) + 20 \times \log(37.04)$$

So,

$$P_T(dB) = 0.87 + 20 \times \log Drj ,$$

where,

$$J/S = 22 \text{ dB}$$

Drj measured in km

This expression is plotted and shown in Figure 64.

In the case of a shipborne jammer, where the same parameters are assumed under optimum propagation conditions, (sea surface attenuation is minimal for both the jammer and the transmitter paths), the same equation stands. The difference here is that the jammer is subject to the surface line of sight propagation conditions, which essentially means that it has to be at distances up to 30 N.M maximum from the intended receiver. This is a severe disadvantage for the jamming platform since it will be virtually within the receiver's SSM range. Therefore, the shipborne jammer is not considered as a likely threat for the system under realistic conditions.

Furthermore, the excessive amount of power required from an airborne jammer to overcome the system's jamming margin is not practically achievable, since there are platform limitations and the ranges are unrealistic and within the receiver ASM range.

2. Smart Jamming

This case includes any possibility, except the 'brute force' technique, that can be used by a potential jammer to overcome the system's high jamming margin.

One of the targets is the system's synchronization channel. This relates to that part of the receiver which recovers the carrier and the local sequence. It includes the carrier local oscillator which is used for phase detection and the local sequence that is used to de-spread the received signal. Three major options can be considered here:

- jamming of the local carrier by CW.
- jamming of the local carrier by noise.
- delay lock loop jamming.

These are further analyzed below.

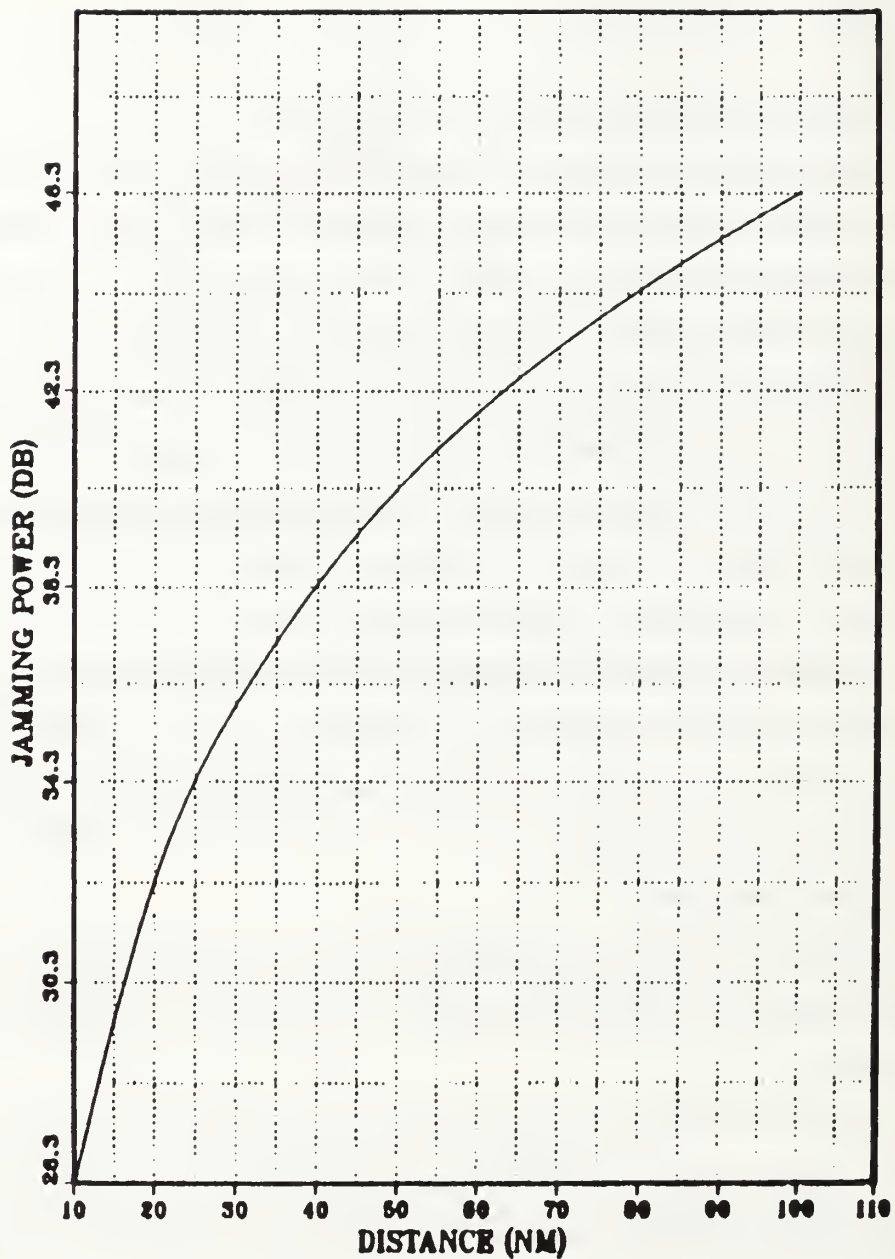


Figure 64. Air-to-Surface Jamming: jamming power vs jammer-to-receiver distance.

a. Jamming of the local carrier by a CW tone

In general, the carrier recovery for a QPSK transmission is achieved with phase lock loops. A strong CW interferer may cause problems

to these circuits resulting in an out of phase and frequency local carrier recovery. A CW jammer is spread and it appears at the loop input of the spread spectrum receiver as noise, thus increasing the RMS tracking error. The configuration of the phase lock loop is shown in Figure 65. A slow change of the jammer carrier (w_j) may cause the VCO to track (w_j) instead of the signal carrier (w_c). The rate of the change of (w_j) is a function of the loop filter bandwidth B_L . The rate of change of the reference frequency is given by the following expression: [Ref. 21].

$$RC = K \times B_L^2$$

where,

RC : rate of change in Hz/sec

B_L : loop bandwidth in Hz

K : a constant typically less than .5

The loop can track frequency changes within this rate, so the jamming signal should have a sweep rate within this limit. This sweep rate must be determined by the jammer, and this is only possible if certain design parameters such as B_L are known.

b. Jamming the local carrier by noise

The noise density of the jamming signal affects the RMS tracking error. This is given by the following formula: [Ref. 22]

$$RMS_{error} = \sqrt{\frac{B_L}{C/No}} \times \sqrt{1 + \frac{W}{2 \times C/No}}$$

where,

RMS_{error} : in radians

W : low pass filter bandwidth or the IF bandwidth

C : the signal power

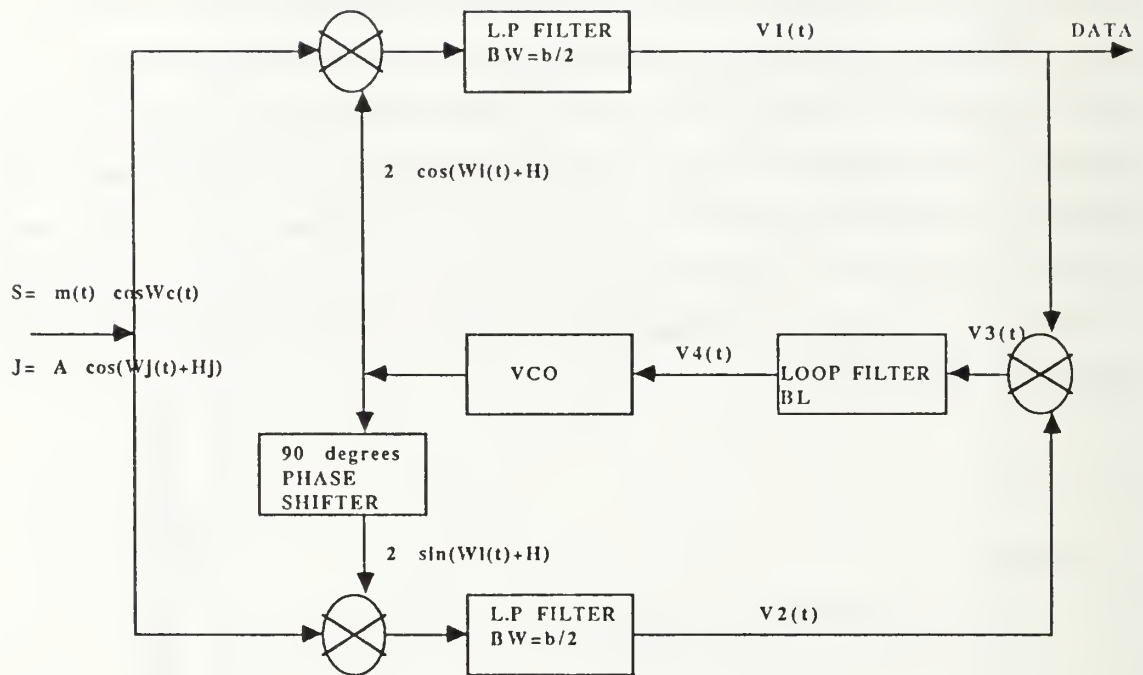


Figure 65. Phase Lock Loop: typical configuration.

The tracking bandwidth B_L is a function of the time required for acquisition and the rate of carrier frequency change due to the Doppler effect caused by the relative Tx-Rx movement. This effect can be expressed by the following formula:

$$F_{\text{doppler-shift}} = \frac{\text{RELATIVE - VELOCITY}}{\text{OPERATING - WAVELENGTH}}$$

If the relative velocity between the TX and the Rx is 100 km/hr and the carrier frequency used is 1200 MHz, the resulting frequency shift is 112 Hz. Then, the minimum required B_L must be around .5 KHz wide to compensate for the positive/negative frequency shifts. Since W is fixed by the information rate used, the lower the B_L the lower the tracking error will be. On the other hand, the B_L cannot go below the minimum specified limit. Also, the B_L must be large enough to allow for fast initial acquisition. The higher the noise power density of the noise jammer the higher is the resulting tracking error. It follows that

certain system parameters must be known in advance by the potential jammer before any effect can be inferred on the tracking error or the required SNR.

c. *Delay Lock Loop Jamming*

Despreading is achieved by having a local reference at the receiver for the generation of the PN sequences. This is done by using delay lock loops (see Figure 66) that help in keeping the local sequence synchronized with the received signal. When there is no phase difference between the received signal and the reference voltage $V_I(t)$, then the code loop filter output voltage is zero and the VCO is set at the normal clock rate. If they are out of phase, an error voltage will result and a correction is made at the clock rate.

Several ways exist to upset this mechanism such as causing the delay lock loop to be out of synchronization (track stealing). This again requires prior knowledge of the Rx-jammer geometry and the sequences used. Also propagation conditions must be known. One option is to adjust the jammer delay in such a way that the jamming signal is within a chip relative to the received signal. If the power of the jammer is high enough to overcome the transmitter's power then the receiver adjusts to the strong jamming. Another option is to change the delay at a suitable rate, thus pulling the tracking loop with the jamming signal and then stop. This leaves the receiver without a signal, forcing it to search again. Synchronization is not possible if the jammer repeats this action. This is depicted in Figure 67. Here, T_o is the time needed to change the receiver AGC, T is the cycle time, Δo is a fixed delay to compensate for the propagation and system delay and Δt is the jamming sequence delay affecting the magnitude of the pulling process. The rate of pulling is given by:

$$RP = .5 \times B_L^2$$

where,

RP: in Hz/sec

B_L : code-loop filter bandwidth

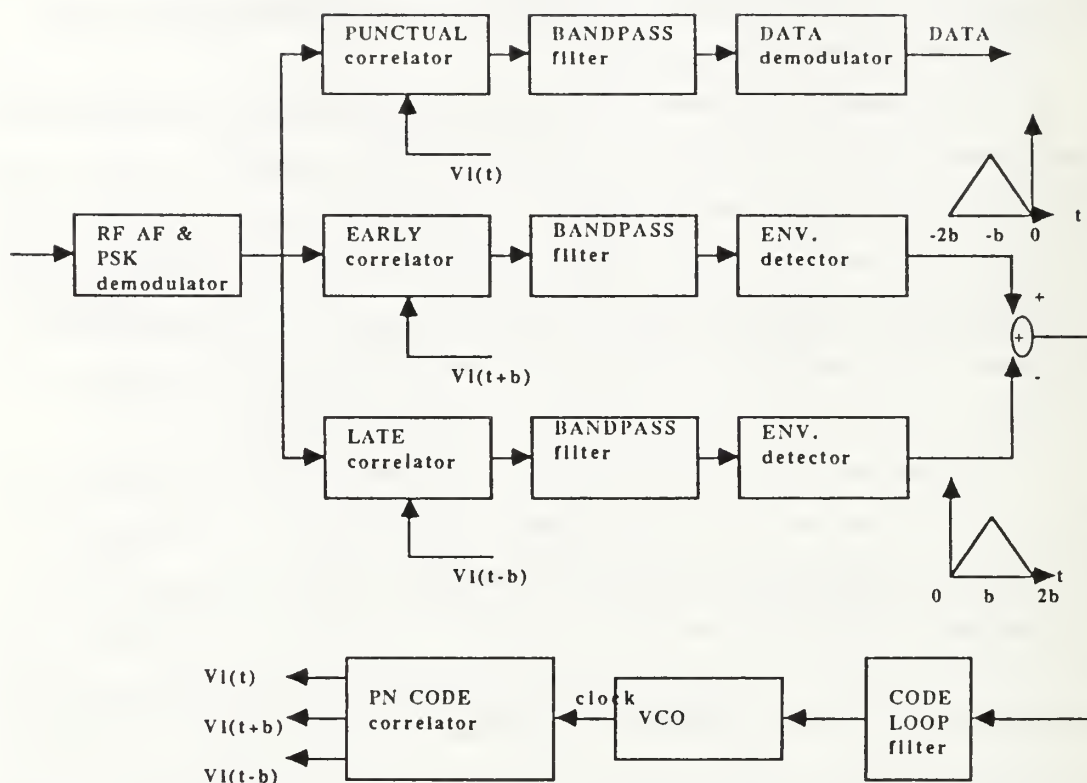


Figure 66. Delay Lock Loop: typical configuration.

The cycle time of pulling (T) can be estimated provided that certain receiver characteristics such as the B_L can be determined by the jammer. The practicality of this technique depends also on the particular geometric situation. Because the jamming signal must be within a chip from the real signal, this means that the jammer must be closer to the intended receiver than the transmitter, when high chip rates are used. In general, the jammer must be at a distance Dr_j :

$$Dr_j = c \times \tau$$

where,

c : speed of light

REPEATER'S
DELAY
TIME

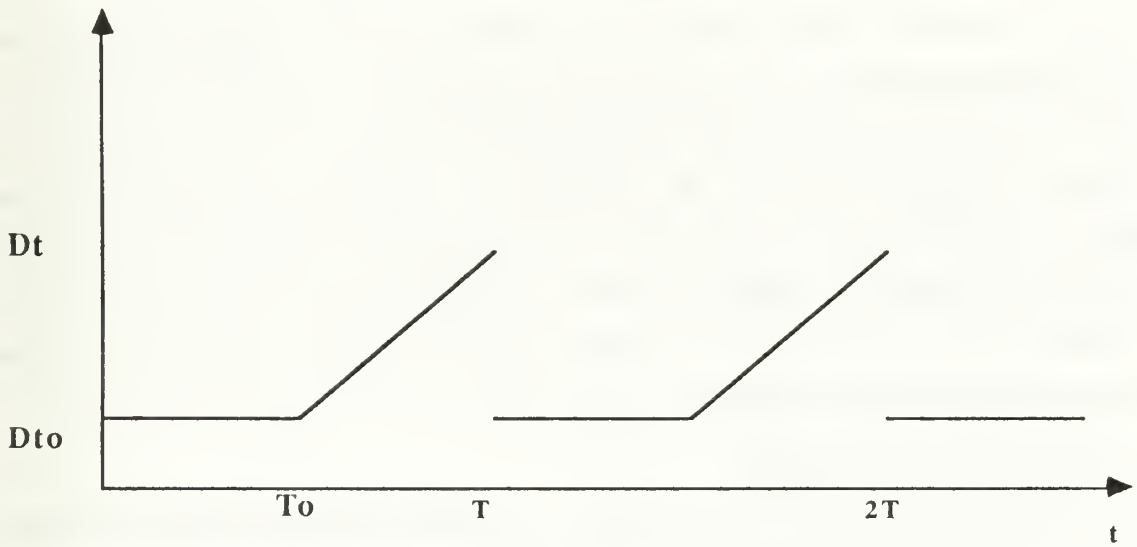


Figure 67. Delay Lock Loop Jamming: pictorial representation.

τ : chip duration

For a system chip rate of 5 Mchips/sec the jammer distance is:

$$Drj = \frac{3 \times 10^5}{5 \times 10^6} = 600 \text{ meters, which is unrealistic}$$

Also, if the jammer doesn't use the same sequence, its power will be spread and it will appear as noise to the delay lock loop. This affects the tracking error as follows: [Ref. 22].

$$\frac{RMS_{error}}{\tau} = \sqrt{B_L \times N_o / (2 \times S)} \times \sqrt{1 + ((2 \times B \times N_o) / S)}$$

where,

B: bandpass filter bandwidth

S: signal power

Again, the jammer needs to throw more and more power (P_J) in order to increase the RMS tracking error. The bandpass filter is fixed by the information rate and the B_L must be at the minimum possible limit to reduce the noise effects. Since the time to regain synchronization after loosing it is inversely proportional to B_L , the desire is to expand it. Optimality can be reached when using an adaptive filter which adjusts its bandwidth in accordance with the situation. Again, the jammer needs to determine the required J/S in order to increase the RMS tracking error. This implies that certain design parameters are already known.

F. SURVIVABILITY ISSUES

A JTIDS-based design is essentially a nodeless system. Therefore the degree of the system survivability is very high because it doesn't require any control nodes for its operation. Communications will be maintained even if only two platforms are left. This design allows for graceful degradation, that is, the loss of any one platform doesn't affect the level of information distribution among the remaining platforms significantly. This platform-to-platform nodeless connectivity enhances the system survivability and it allows communications reconstructability in the event of several platform losses.

V. LIFE CYCLE COST ANALYSIS

A. BACKGROUND

Life cycle costing is a systematic analytical process of determining and listing the total cost of developing, producing, owning, operating, supporting and disposing of equipment or complete systems. A major objective of life cycle costing is to provide decision makers at all levels with sufficient economic information to determine the most cost-effective configuration for a system within budget limitations [Ref. 23: p.4].

Several items must be considered when evaluating project costs. These are:

- The cost breakdown structure.
- The Life Cycle Cost (LCC).
- Discount Rate and Learning Curve.
- Risk analysis.

The project work breakdown structure provides a format for identifying and organizing detailed cost information. The LCC represents the program cost throughout the life of the project. The discount rate and any existing learning curves provide correction figures for the time value of money, when analyzing life cycle costs that occur in different periods, and for cost changes that occur with production experience. Since there is uncertainty about what costs occur, when and in what amounts, some method of assigning probabilities to these costs using risk analysis must also be employed.

The estimation of several cost elements is based on the development of relevant cost relations which in turn compose a cost model.

A parametric or a statistical Cost Estimating Relation (CER) can be derived for new systems if there is historical data from prior systems that are functionally similar. Once a CER is derived, it can then be used to estimate the costs associated with the new system by direct substitution of the various design parameters and performance specifications into the cost-equation. And

this can be done for all the alternatives [Ref. 23: p.24]. For example, a typical cost estimate is derived as follows:

$$\text{Cost} = aX + bY + cZ + \dots$$

Where, X,Y,Z.... : are cost factors (design parameters and performance specifications) and a, b, c.... are CER coefficients based on the analysis of historical data of the appropriate equipment.

This cost-model is particularly useful during the system design stage where only mission and performance envelopes are defined.

B. METHODOLOGY FOR LIFE CYCLE COST ANALYSIS

The methodology chosen here is the one used for the evaluation of joint tactical communications program costs (TRI-TAC methodology). This includes eight steps as shown in Figure 68. These steps are closely followed below for the actual system cost analysis [Ref. 23: p.29].

1. Objectives

The purpose of this analysis is to determine if a JTIDS-based system is an economically acceptable solution when budget limitations exist. In this sense, it is an effort to evaluate a total system cost figure which in turn can be used to assist the decision maker in realizing the magnitude of the investment required for the procurement of a JTIDS- based system in current dollar figures.

This analysis recognizes that there is no alternative design, available in the open market, that approaches the effectiveness of a JTIDS-based design in the area of joint tactical communications systems. The lack of comparable alternative designs eliminates the possibility of competitive procurement and precludes a cost-effectiveness analysis of such a system.

Considering these limitations, this analysis aims to reach economic conclusions that can help in assessing future decisions concerning investments on such systems.

2. Assumptions

Several assumptions are considered pertinent to this analysis. These are broken down in two categories. The first category concerns the develop-

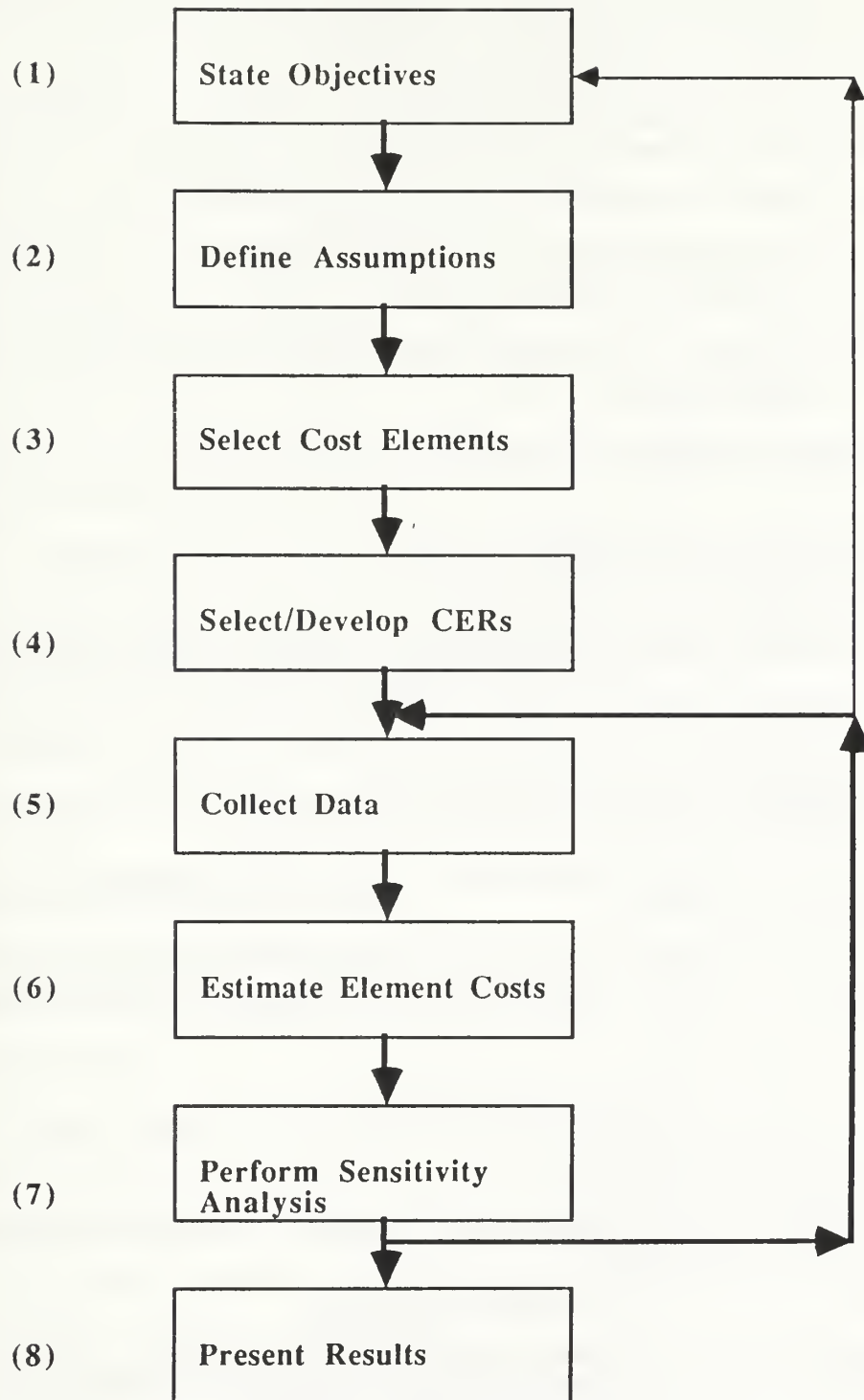


Figure 68. The LCC Methodology: The eight steps for cost analysis [Ref. 23: p.30].

ment of an ideal cost-model for the system given that there are no limitations in obtaining the actual detailed data required for using this model. The ideal model is proposed here to demonstrate the actual way of performing the analysis for this system. The second category of assumptions is considered necessary because the ideal life cycle cost model has to be modified in this analysis. This is mainly due to the limited availability of JTIDS-related historical data below the terminal level, and the confidentiality of detailed information about the specific cost elements. These assumptions are stated below:

a. First Category

- Approval for official release of the system to specific foreign customers is assumed here, and the details of this are beyond the scope of this study.
- The number of different classes of terminals required for system deployment is arbitrary but realistic.
- The total number of various platforms is arbitrary but realistic (Eighty total: seventy shipborne and airborne versions and ten ground versions).
- The number of sites and bases at which these platforms are deployed is chosen arbitrarily.
- The number of different replaceable items within each terminal is chosen from existing historical data about similar systems.
- The number of support type items is chosen from existing historical data about similar systems.
- The monetary figures (unit prices) are chosen from existing historical data about similar systems. They are not actual values but they are realistic.
- Inflation is ignored.
- Salvage cost is assumed to be negligible.

b. Second Category

- Operation and Support costs are not considered here due to the absence of relevant historical data.
- Total service time for the system is 20 years.
- Aggregate figures for Research and Development (R&D) and Production costs are based on historical data about systems similar to JTIDS. These data were provided on a system-terminal basis [Ref. 24].
- The R&D costs for the actual system are assumed to constitute either a fair portion of the JTIDS R&D costs, when the system is procured through FMS (Foreign Military Sales) channels, or they are assumed to be negli-

gible when the system is procured by directly contacting the prime manufacturer. The latter also implies the existence of an open competitive market for such systems. There is a middle case, where R&D costs can be somewhere in between the previous two alternatives. The extreme case of procuring a totally new system is not considered, since it is unacceptable due to the extraordinary R&D-to-Investment cost ratio for a project with small-scale production requirements.

- The learning curve slope is derived from the commonly acceptable range of typical slopes for communications and electronic systems. These are the most frequently used parameters within the cost community. Here the learning curve slope is chosen to be 92 percent over a period of 6 years (1988-1993). Production costs after this period are assumed to follow the actual learning curve observed in the historical data [Ref. 23: p.37].
- The discount rate used is 10 percent which is the current DOD conventional discount rate [Ref. 23: p.47].
- R&D modification costs for installation on maritime patrol aircraft (MPAs) are considered to be 3 percent more than the standard airborne version. There is no difference between the costs of the JTIDS command and control airborne (E2-Cs) terminal version and the respective ground command and control terminals for the proposed system.

3. Selection of cost elements and proposed LCC model development

The system's LCC structure breaks down into four aggregate areas (see Figure 69).

- Research and Development costs.
- Production and Investment costs.
- Operation and support costs (provisional time).
- Salvage costs.

Because salvage costs are considered to be negligible, this area will be eliminated from this discussion. The remaining three aggregate areas are further broken down into fifteen (15) distinct cost elements (CE), that constitute the System LCC model. These elements together with their respective aggregate areas are shown in Figure 70. [Ref. 25].

This model accepts input from both the customer and the contractor. The customer inputs include operational and environmental information, the number of systems that will be online, and the number of hours the system will operate. The contractor's inputs include the mean time between failure

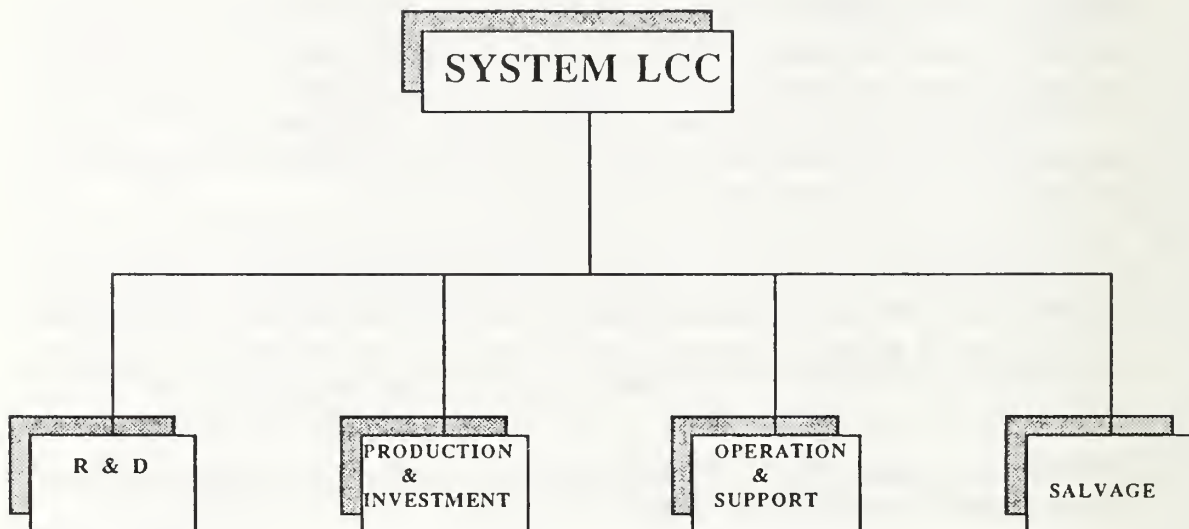


Figure 69. The LCC Structure: aggregate areas.

(MBTF), the maintenance cycle of the equipment and the overhaul requirements of the system. The three aggregate areas constitute the whole system LCC for all terminals in all platforms at all sites and bases where the system is installed.

4. Select/Develop CERs

Each cost element in Figure 70 can be expressed through a parametric or a statistical CER which in turn may be used to evaluate the actual cost. These CERs are described and expressed below. These CERs presume that data is available for the ideal model, thus these CERs reflect the first category of assumptions. Specific notation used for these expressions is shown in Table 10.

a. Description of CERs for Ideal LCC Model

(1) *CE1: Hardware Acquisition.* This element refers to the acquisition cost of primary mission equipment, namely the system terminals and the integration equipment required to interface with the platforms. This can be expressed through a parametric CER:

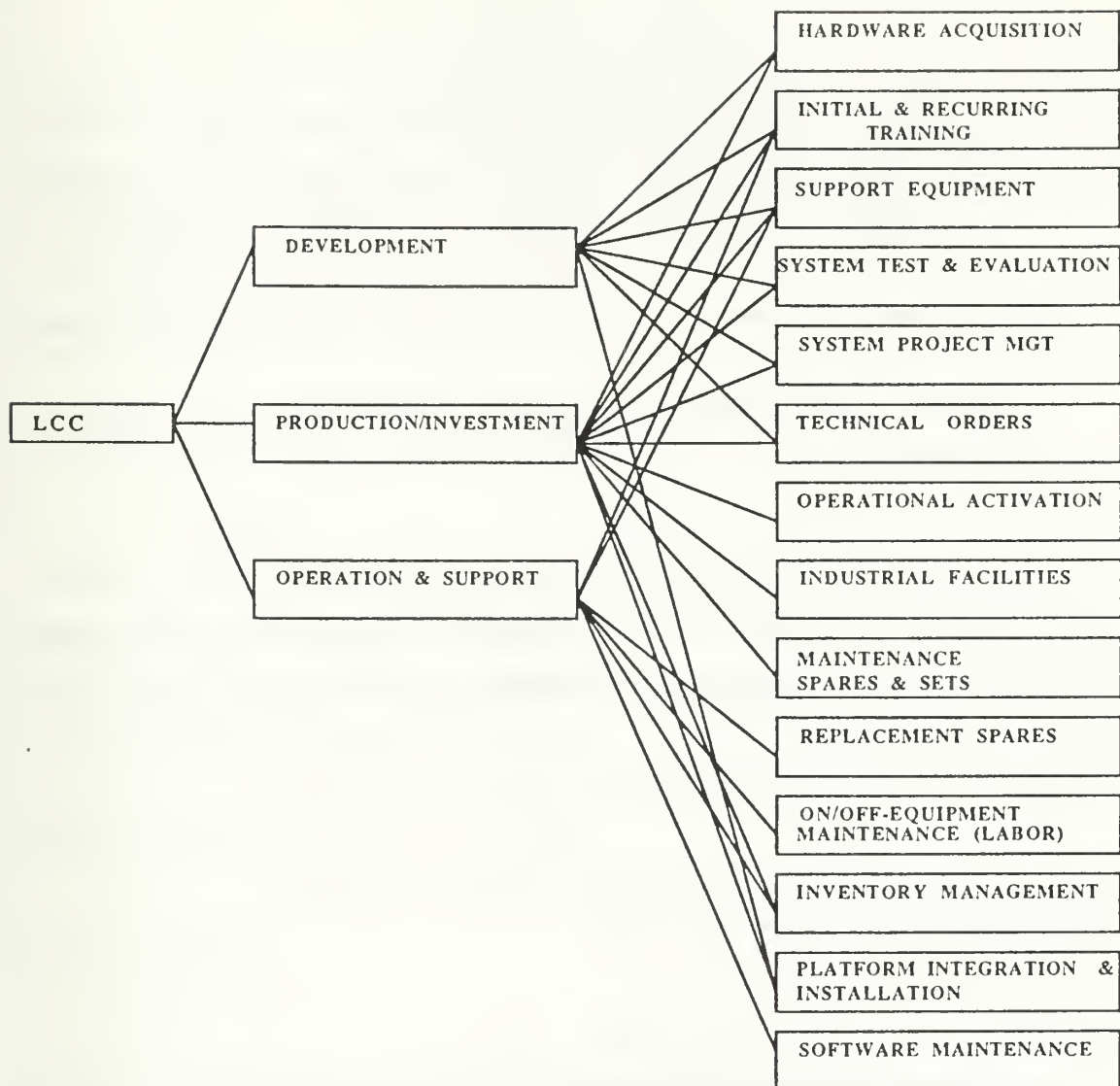


Figure 70. The LCC Model: The fifteen system cost elements.

$$CE1 = \sum_I K(I) \times UP(I)$$

save Where,

K(I): number of terminals of class (I) to be installed in all system platforms of similar type. (Terminal # and class) (Customer)

Table 10. GENERAL INDEX NOTATION FOR CERS

I	K	NP	ICM	NS	L
Item Type	Terminal Class	Platform Type	Command Type	Specific Bases	Support Equipment Item Type

UP(I): unit price for terminal of class (I). (\$/Item)
(Customer/Contractor)

(2) *CE2: Spares Acquisition.* This element consists of the initial investment cost for item spares required to support the bases and it includes inventory costs at each base. This can be expressed through a parametric CER :

$$CE2 = \sum_I \sum_{NS} [SB(I,NS) \times NB(NS)] \times UP(I)$$

Where,

SB(I,NS): number of spare items of terminal class (I) to be acquired for inventory at base NS. (Items) (Customer/Contractor)

NB(NS): total number of platforms deployed at base type NS. (Sites) (Customer)

(3) *CE3: Maintenance Sets.* This element refers to the acquisition cost of maintenance sets. These sets are to be used for on-equipment item maintenance. This can be evaluated through a statistical CER based on historical data. (Customer/Contractor)

(4) *CE4: Off-Equipment Maintenance.* This element refers to base-level repairs performed on items which must be removed from their respective terminals. This includes the labor cost incurred from time to diagnose, repair or attempt to repair. This can be evaluated through a statistical CER based on historical data. (Contractor/Customer)

This cost is mainly driven by the monthly failure rate of item class I in terminal type K at base NS. This is shown with the following CER:

$$Fail(I,K,NS) = \frac{NITEM(I,K) \times NPLT(K(I),NS) \times APFH(K(I))}{MBTF(I)}$$

Where.

NITEM(I,K): number of items class I in terminal type K installed on a platform class I. (Contractor)

NPLT(K(I),NS): number of platforms carrying terminals class I deployed at base type NS. (Customer)

APFH(K(I)): average monthly operating hours for platform carrying terminals of class I. (Customer)

MBTF(I): Mean time between failures for an item class I. (Hrs) (Contractor)

(5) *CE5: Replacement Spares.* This element represents the LCC of terminal replacement items. It includes the costs of repairable items purchased to replace those items which are designated for discard upon failure and those that are not repairable. This can be also evaluated by a statistical CER based on monthly failure rate (see CE4). (Customer/Contractor)

(6) *CE6: On-Equipment Maintenance.* This element refers to the LCC of all work performed on items at their respective terminals. It can be estimated by a statistical CER driven by the monthly failure rate (see CE4). (Customer/Contractor)

(7) *CE7: Support Equipment.* This refers to the LCC of system support equipment purchased for each maintenance facility (base). An additional fraction of the cost is included to allow for the maintenance of the support equipment, including both labor and spare parts, over the system life cycle. It can be calculated using historical data or through the following parametric equation:

$$CE7 = \sum_L \sum_{NS} [NAPB(L,NS)] \times UP(L) \times (1 + PIUP \times AMA(L))$$

Where,

NAPB(L,NS): number of support items of class L purchased for base NS. (Items) (Customer/Contractor)

UP(L): unit price of the item support equipment of class L. (\$/Item) (Contractor/Customer)

PIUP: Program Inventory Usage Period (operational service life of the system). (Yrs) (Customer)

AMA(L): average annual portion of the fraction of the acquisition cost of item support equipment of class L regarded as maintenance cost of this item over the life cycle. (Decimal Fraction/Yr) (Contractor)

(8) *CE8: Initial Training.* This refers to the costs needed for training the initial maintenance and operator/specialist personnel, and the cost of specific training equipment required for the system. It can be estimated based on historical data, through a statistical CER.

$$CE8 = ITEC + IOPC + IBMC$$

Where,

ITEC: cost of required training equipment. (\$) (Customer/Contractor)

IMPC: cost of initial training for on-equipment maintenance and operator/specialist personnel. (\$) (Customer)

IBMCM: cost of initial training for off-equipment maintenance personnel. (\$) (Contractor/Customer)

(9) *CE9: Recurring Training.* This refers to the cost required to train the maintenance and operator/specialist personnel needed to replace the initially trained personnel who have been transferred, promoted or have left the armed services. It can be calculated using historical data. (\$/Man/Week or \$/Man) (Customer/Contractor)

(10) *CE10: Inventory Management.* This refers to the costs required to introduce new assemblies and parts into the customer inventory system, together with the recurring supply inventory management costs associated with such inventories. It can be estimated using historical data. (\$/Item/Yr) (Customer)

(11) *CE11: Technical Orders.* This refers to the cost of purchasing the master negatives for technical orders, overhaul manuals and other repair documentation. (\$/Page) (Contractor/Customer)

(12) *CE12: Full Scale Development.* This refers to the cost of developing primary mission equipment, support equipment and associated software. This is an aggregate cost of the basic inputs. It can be computed both parametrically or statistically.

$$CE12 = \sum_I DEVL(I) + \sum_K \sum_H DEVT(K,H) + \sum_L \sum_H DEVSE(L,H) + \sum_H DEVSYS(H) + DEVSAL + DEVBMS$$

Where,

DEVL(I): development cost for item class I. (\$/Item)
(Customer/Contractor)

DEVT(K,H): dummy variable for development cost for terminal class K, for hardware (H = 1) or software (H = 0). (\$/Terminal) (Customer/Contractor)

DEVSE(L,H): dummy variable for development cost for support equipment item class L, for hardware (H = 1) or software (H = 0). (\$/Support Item) (Customer/Contractor)

DEVSYS(H): development cost not related to the terminals. (\$/System) (Customer/Contractor)

DESVAL: cost of engineering design and validation of the system. (\$) (Contractor)

DEVBMS: development cost of maintenance sets. (\$) (Contractor)

(13) *CE13: Platform Integration.* This refers to the cost needed to develop the integration hardware and software required to interface the system terminals to existing platform equipment. It can be calculated or estimated both parametrically or statistically.

$$CE13 = \sum_I \sum_{NP} DEVL(I)$$

Where,

DEVL(I): development cost associated with those items that are used as an interface between some terminals and some platform. (\$/Item)
(Customer/Contractor)

(14) *CE14: Platform Installation.* This refers to the cost of installing terminal and integration items, together with the cost to purchase items classified as racks, cables, and similar equipment. It can be estimated or calculated both statistically or parametrically.

$$CE14 = \sum_I \sum_{NS} \sum_{NP} [WT(I) \times CERSTS(NP) \times DLR + UP(I) \times NINSTL] \times \sum_K \sum_{ICM} NITEM(I,K,NP) \times MPS(NS,NP,ICM) \times NB(NS)$$

Where,

WT(I): item weight. (Lbs) (Contractor)

CERSTS(NP): man hours per pound to install an item in platform type NP. (Man-Hrs/Lb) (Customer/Contractor)

DLR: labor rate. (\$/Man-Hr) (Customer/Contractor)

UP(I): average unit cost of racks, cables and similar equipment. (\$) (Contractor)

NINSTL: number of installation items required per item (\$) (Contractor)

NITEM(I,K,NP): quantity of items type I in terminal class K installed in platform of type NP (Items) (Contractor/Customer)

MPS(NS,NP,ICM): number of platforms of type NP that are located at site NS and utilized by command ICM. (Platforms) (Customer)

NB(NS): total number of sites of type NS (Sites) (Customer)

(15) *CE15: Software Maintenance.* This refers to the cost required to support all system associated software, both in terminals and in the host computers at the platforms in which the system is installed. It can be estimated using historical data. (\$) (Customer/Contractor)

b. Modified Cost Model

The previously derived LCC model can be used provided that data is available. But the lack of detailed historical data led to the use of a modified cost model that closely follows the second category of assumptions made in the beginning of this chapter.

On the basis of the previously analyzed cost elements, the work break-down structure and the aggregate figures available for these terminal versions, two of the major cost elements have been calculated using the following relations: [Ref. 24].

Research, Development and Evaluation (RDE):

$$RDE(t) = R_{FSD} + R_{PLI} + R_{GIH} + R_{TEV}$$

Where,

R_{FSD} : contract actuals on full scale development

R_{PLI} : platform integration historical costs

R_{GIH} : government-in-house historical RDE costs

R_{TEV} : test and evaluation historical costs

Data for these elements was actually provided in the following form:

$$RDE(t) = R_{FSD/PME} + R_{PLI} + R_{TEV} + R_{ILS/ENG}$$

Where,

$R_{FSD/PME}$: cost of FSD of prime mission equipment

$R_{ILS,ENG}$: Development cost for engineering effort, logistics and support.

Production and Investment (I):

$$I(t) = I_{HW} + I_{GIH} + I_{IS} + I_{ILS} + I_{ECP} + I_{NS}$$

Where,

I_{HW} : FSD hardware adjusted for production

I_{GIH} : government-in-house (12% of hardware costs)

I_{IS} : initial spares cost equivalent to hardware

I_{ILS} : initial logistics & support cost (11% of flyaway)

I_{ECP} : 5% of hardware spares

I_{INS} : installation costs (for Aircraft: 5%, for ships: 20% of hardware)

Data for these elements was actually provided in the following form:

$$I(t) = I_{PME} + I_{GIH} + I_{ILS}$$

Where,

I_{PME} : production cost for prime mission equipment

5. Collect Data

A matrix approach has been used to organize the cost data for these two cost elements. The process is illustrated in Tables 11 and 12 [Ref. 26: p.23-33 and 36-37].

Table 11. SYSTEM RDE COST ESTIMATE FOR 792 TERMINALS (\$M)

COST ELEMENT	FY88	FY89	FY90	FY91	FY92	FY93	TOTAL
FSD PME	23.13	40.71	49.15	37.13	21.40	22.89	194.3
Platform Integration	39.21	60.63	70.43	58.80	31.54	33.07	293.68
Test and Evaluation	22.27	22.57	26.78	27.41	27.71	48.10	174.87
System Eng. and ILS	16.80	23.56	29.70	23.30	21.65	50.06	165.16
RDE TOTAL	101.41	147.33	176.14	146.66	102.68	154.15	828.37

The figures for platform integration include a 3% increase for the modification needed for the MPA terminal. No further changes have been incorporated here since all the other terminal versions follow the JTIDS-based design figures.

Table 12. SYSTEM PRODUCTION COST ESTIMATE FOR 792 TERMINALS (\$M)

COST ELEMENT	FY88
Prime Mission Equipment	497
Government-in-House	86
Installation	99
TOTAL	682

The installation and production costs (Table 12) include related figures for the airborne versions (FAC and MPA), and the shipborne and ground versions as well.

The aggregate LCC can be computed from the following relationship:

$$LCC = \sum_t \frac{1}{(1+r)^t} \times [(R(t) + I(t) + S(t) + Salvage)]$$

Where,

r: discount rate

R(t): aggregate R&D costs during time t

I(t): aggregate investment and production costs during time t

S(t): aggregate support costs during time t

According to the assumptions listed earlier, cost data has been collected for R(t) and I(t). S(t) and salvage value have been ignored. The data has been provided by the Naval Center for Cost Analysis and reflect JTIDS cost figures. Furthermore, the cost data should be discounted to account for the time value of money and procurement costs should be adjusted for learning effects.

The rationale behind discounting is the fact that there is a time preference in consumption such that a present monetary unit is worth more than a

future monetary unit. Then it follows that procurement of that unit in the present represents a greater expenditure value than procurement of it in the future. Consequently, discounting should be considered when adding together dollars spent or received in different periods, because they have different values

The problem then lies in the proper choice of the discount rate. Ideally, in a market economy, if there weren't any taxes or other kinds of distortions, there would be a unique discount rate and this would be the market rate of interest which would be the same as the consumer rate of interest. Since this is not the case, the discount rate should optimally reflect the ratio of consumer willingness to consume today versus the willingness to consume tomorrow. This optimality is determined at the point where the production possibility frontier curve and the social indifference curve are tangent to one another. This then becomes the social rate of time preference. But since this is something that cannot easily be measured, several approximations have been suggested. The opportunity cost rate, that normally reflects a government's opportunity cost of capital appears to be a good approximation for a proper discount rate [Ref. 27]. But for the purpose of this analysis, and given that data for the exact measurement of this is not available, a 10% discount rate is used. This is the official discount rate for evaluating U.S government projects [Ref. 28]. However, when evaluating time costs, the analyst must keep in mind these lines, for a proper and accurate estimation [Ref. 23: p.47-49]. (Note that all costs have been expressed in real terms so corrections for inflation are unnecessary).

The cost also needs to be corrected using existing learning curves. The learning curve correction is based on historical evidence that as the total quantity of units increases, the manhours or cost to produce this quantity decreases by some constant rate. The common measure of the learning slope is expressed as the ratio of change in cost when the quantity is doubled. Figure 71 shows the typical learning curve [Ref. 23: p.37-43] and [Ref. 29].

The cumulative average aggregate investment (production) cost (AVI_N) of N items is given by the following formula:

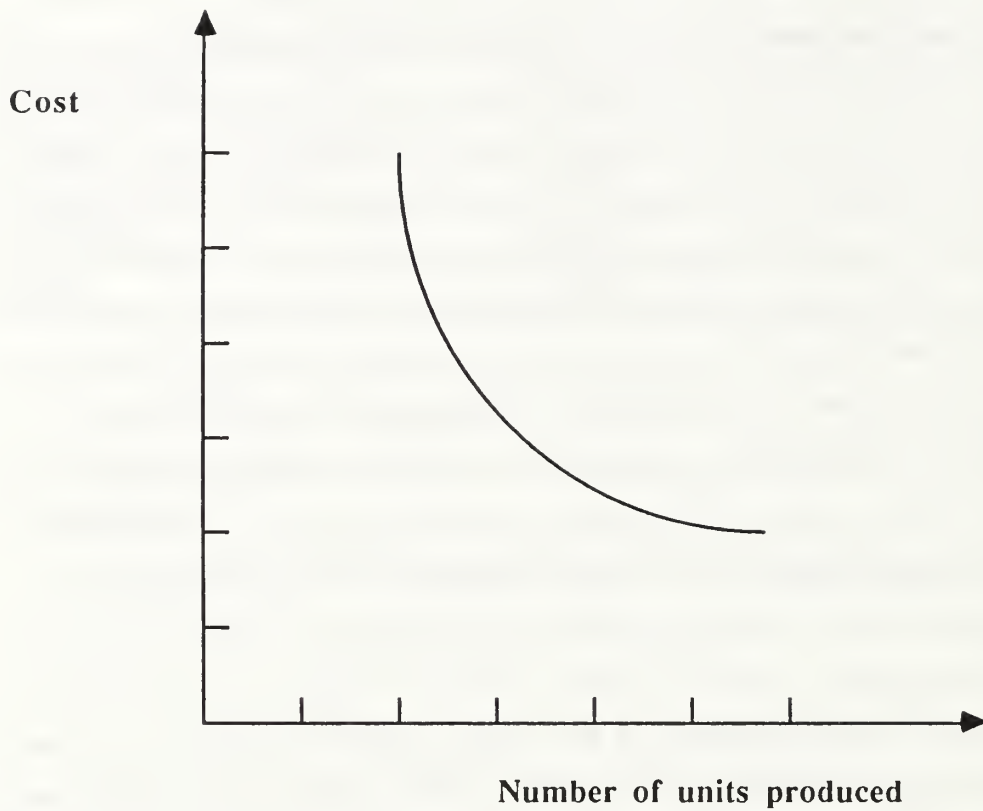


Figure 71. Typical Learning Curve: exponential decline.

$$AVI_N = A \times N^B$$

Where,

N: number of units produced in production lot

A: theoretical cost of first unit in production lot

B: correction slope factor

$$B = \log S / \log 2$$

and,

S: slope

The investment (production) cost of the individual N^{th} unit from a particular lot of units ordered is given below:

$$I_N = A \times [N^{(1+B)} - ((N - 1)^{(1+B)})]$$

Then the total investment (production) cost Tl_N of the lot can be expressed as:

$$Tl_N = A \times (N^{(1-B)})$$

The production cost data in this analysis has been adjusted for learning curve effects based on the procurement plan shown in Table 13.

Table 13. SYSTEM TERMINAL PROCUREMENT PLAN

TERMINALS	FY94	FY95	FY96	FY97	FY98	TOTAL
Ground	1	1	2	2	2	8
Ground Trainers	1	1				2
Fighter AC	4	4	4	4	4	20
FAC Trainer	1	1				2
MPA	2	2	2	2	2	10
MPA Trainer	1	1				2
Ships	6	7	7	7	6	33
Ship Trainer	1	2				3
Grand Total						80

6. Estimate Element Costs

a. RDE Costs

Based on the assumptions made for the RD&E cost, we get three different cost estimates as below:

(1) *Case 1: Procurement through FMS.* In this case the RDE cost is a function of the number of terminals ordered by the U.S, the number of terminals ordered by all non U.S countries, and the number of terminals ordered by the customer under consideration. The allocation of the RDE cost is based on an algorithm which splits it into a 'fair' proportion for each country. This algorithm, as shown below, is a pure averaging formula:

$$RDE_{FMS} = \left(\frac{RDE}{X + Y} \right) \times N$$

Where,

X: number of terminals ordered by the U.S

Y: number of terminals ordered by non U.S countries

N: number of terminals ordered by the particular country

Using the historical data from Table 12 and a discount rate of 10 percent, the discounted R&D cost for seven hundred ninety two (792) JTIDS-based terminals is:

$$RDE = \sum_{t=1}^6 \frac{R(t)}{(1+r)^t} = 597.23\$M$$

Then by applying the 'fair' proportion algorithm in allocating R&D costs through FMS, the actual system's RDE cost for eighty (80) terminals becomes:

$$RDE_{FMS} = \left(\frac{597.23}{(792 + 500 + 80)} \right) \times 80 = 34.82\$M$$

Where, X= 792, Y= 500+80, N= 80, for a total lot of 1372 terminals.

(It is assumed that the U.S. has ordered 792 terminals and that 500 terminals have been ordered by other non U.S countries prior to the order placed by the country under consideration).

(2) *Case2: Procurement through the Prime Manufacturer and indirect use of Economic Aid tied to FMS.* Here the RDE cost is assumed to be 5% of the total RDE cost incurred for the development of the 792 JTIDS U.S terminals. This slightly lower figure reflects the use of part of the economic aid provided to the country under consideration and tied to FMS, to offset a portion of the RDE costs allocated to that country. The resulting figure is:

$$RDE_{FMSPM} = 5 \% \text{ of } RDE = 29.86\$M$$

(3) *Case3: Procurement through the direct contact with the Prime Manufacturer.* In this case the RD&E costs are zero, assuming a perfectly competitive market for the particular system.

$$RDE_{PM} = 0$$

b. Production and Investment Costs

For production costs, the data were adjusted for a learning curve slope correction, based on the previous analysis and the relevant assumptions. The adjustment was made as follows:

$$B = \frac{\log .92}{\log 2} = -.12$$

$$AVI_N = \frac{682}{792} = .86\$M$$

$$A = \frac{(AVI_N)}{(N^B)} = \frac{.86}{.448} = 1.91\$M$$

$$I_{1292} = 1.91 \times [(1292^{.88}) - (1291^{.88})] = .70\$M$$

$$I_{1372} = 1.91 \times [(1372^{.88}) - (1371^{.88})] = .70\$M$$

$$TI_{1292} = .70 \times (1292^{.88}) = 386.49\$M$$

$$TI_{1372} = .70 \times (1372^{.88}) = 407.48\$M$$

$$TI_{80} = TI_{1372} - TI_{1292} = 20.99\$M$$

$$(AVI_{80}) = \frac{20.99}{80} = .262\$M$$

Production cost data has already been discounted, so no further adjustment is required.

c. Total System Cost Figures

The cost figures for the proposed system, discounted and adjusted for learning effects are shown in Table 14. These calculations have been performed assuming a JTIDS-based design.

Table 14. SYSTEM AGGREGATE COST ESTIMATE FOR 80 TERMINALS, DISCOUNTED AND ADJUSTED FOR LEARNING EFFECTS (\$M)

PROCUREMENT METHOD	TOTAL PROJECT COST	AVERAGE COST/TERMINAL
FMS	$34.82 + 20.99 = 55.81$.697
FMS & Prime Manufacturer	$29.86 + 20.99 = 50.85$.635
Prime Manufacturer	20.99	.262

7. Sensitivity Analysis

The results presented previously are based on historical cost figures, a particular discount rate and a specific learning curve slope. Consequently, there are three sources of uncertainty associated with the previous analysis. The first is uncertainty about the cost figures used, because these are not the actual but rather historical costs expected to reflect the actual costs [Ref. 25]. Second is the subjective choice of a discount rate, and third is the use of an estimated learning curve slope. Sensitivity analysis is appropriate to understand the impact of these uncertainties on the LCC estimates.

Concerning the cost data, it is more meaningful to assign probabilities of attainment to them rather than only comparing their central tendency values. This probabilistic cost analysis helps in evaluating the degree of uncertainty surrounding a most probable cost value. In this case, the system cost elements are not assigned specific values but rather they are characterized by their cost probability distributions. This method requires the determination of these probability densities.

Concerning the discount rate and the learning curve slope, it is important to determine the degree of cost sensitivity (or no sensitivity) when these two correction factors vary over a wide range of values. This analysis is performed and explained in the following sections.

a. Sensitivity analysis for the cost figures

This is based on two assumptions:

- The use of the triangular distribution for the various cost elements is the most pertinent since it requires the least amount of information for a meaningful description of cost estimation. It requires only three parameters for each cost: the minimum possible cost (a), the maximum possible cost (b) and the most probable cost (m). This is shown in Figure 72 [Ref. 26 p.39-42].
- Actual costs are assumed to range between 20% above and 5% below the most probable cost (m), which in this case represents the JTIDS historical figures. Normally the most probable cost (m) is the average of the maximum possible cost estimate (b) and the minimum cost estimate (a) over the range of possible values that can be obtained [Ref. 26 p.39-42].

Based on these assumptions, the expected value $E(c_i)$ and the variance $V(c_i)$ for the i^{th} cost element can be computed by using the following formulas:

$$E(c_i) = \left(\frac{1}{3 \times (b - a)} \right) \times \left[\left(\frac{m^2 \times (2m - 3a) + a^3}{(m - a)} \right) + \left(\frac{m^2 \times (2m - 3b) + b^3}{(b - m)} \right) \right]$$

$$V(c_i) = E((c_i)^2) - [E(c_i)]^2$$

The results of this analysis are given in Table 15 below.

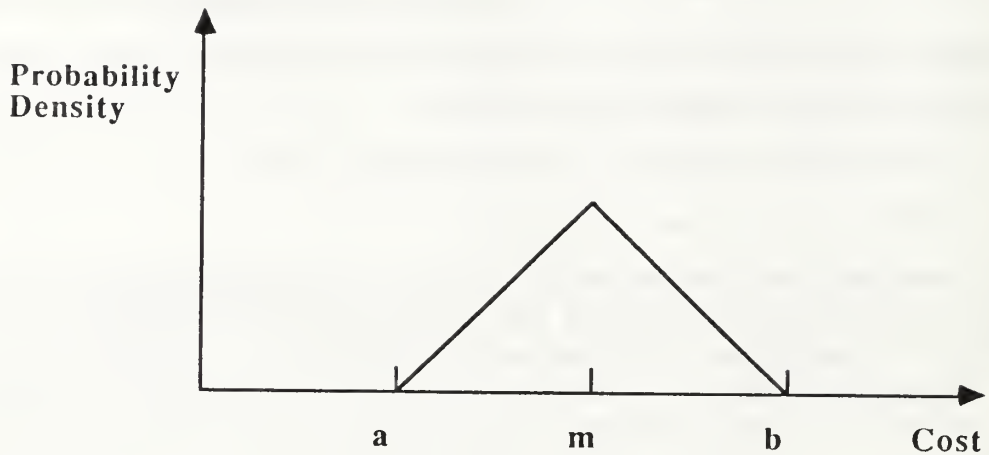


Figure 72. Cost Estimation Description: triangular cost probability distribution.

Table 15. RDE COST ESTIMATE RANGE DETERMINATION FOR 792 TERMINALS (FY88) (\$M)

COST ELEMENT	ESTIMATION RISK (DOWN)	ESTIMATION RISK (UP)	HISTORICAL ESTIMATE (m)	LOWER LIMIT (a)	UPPER LIMIT (b)	MEAN $E(c_i)$
PME	-1.15	+ 4.62	23.13	21.98	27.75	24.2
Platform Integration	-1.96	+ 7.84	39.21	37.24	47.04	40.85
Test and Evaluation	-1.11	+ 4.45	22.27	21.16	26.72	23.24
System Eng and ILS	-.84	+ 3.36	16.801	15.96	20.16	17.64
OVERALL MEAN/STD DEVIATION						106.48/8.36

Basically five more tables of the same nature have been analyzed for the whole period (1988-1993). The risk assessment for the aggregate RDE cost estimate and range determination is shown in Table 16 below.

Table 16. TOTAL RDE COST ESTIMATE RANGE DETERMINATION FOR 792 TERMINALS (\$M)

FIS- CAL YEAR	ESTI- MATION RISK (DOWN)	ESTI- MATION RISK (UP)	HIS- TORICAL ESTI- MATE (m)	LOWER LIMIT (a)	UPPER LIMIT (b)	OVER- ALL MEAN $E(c_i)$	OVER- ALL STDEV
FY88	-5.07	+20.28	101.41	96.34	121.69	106.48	8.36
FY89	-7.36	+29.46	147.33	139.97	176.79	154.93	15.49
FY90	-8.8	+35.22	176.14	167.34	211.36	184.92	17.5
FY91	-7.33	+29.33	146.66	139.33	175.99	153.95	13.73
FY92	-5.13	+20.53	102.687	97.55	123.217	107.8	4.25
FY93	-7.7	+30.83	154.15	146.45	184.98	161.76	11.16

By the central limit theorem we have that the sum of mutually independent random variables with a common distribution will be asymptotically normally distributed with a mean and variance of :

$$E(c) = \sum_{i=1}^n E(c_i)$$

$$V(c) = \sum_{i=1}^n E(c_i^2) - [E(c_i)]^2$$

So the overall range for each year is:

$$\text{Range: } [E(c) - 2\sigma, E(c) + 2\sigma]$$

The results are given in Tables 17 and 18 for both the RDE and the production costs below:

And after discounting with a factor of 10 percent the RDE cost range becomes:

RDE cost range (\$M): (522.8, 730.86)

Table 17. TOTAL RDE COST ESTIMATE RANGE DETERMINATION FOR 792 TERMINALS (\$M)

FISCAL YEAR	88	89	90	91	92	93
RANGE	(89.28, 123.2)	(123.95, 185.91)	(149.92, 219.92)	(126.49, 181.41)	(99.3, 116.3)	(139.44, 184.08)

Also, the overall range for the production cost is:

Production cost range (\$M): = (516.09, 916.09).

Table 18. PRODUCTION COST ESTIMATE AND RANGE DETERMINATION FOR 792 TERMINALS (\$M)

COST ELEMENT	ESTI-MATION RISK (DOWN)	ESTI-MATION RISK (UP)	HIS-TORICAL ESTI-MATE (m)	LOWER LIMIT (a)	UPPER LIMIT (b)	MEAN $E(c_i)$
PME	-24.85	+ 99.4	497	472.15	596.4	521.8
Government in-House	-4.3	+ 17.2	86	81.7	103.2	90.28
Installation	-4.95	+ 19.8	99	94.05	118.8	103.88
OVERALL MEAN/STD DEVIATION						716.09/200

The figures then are converted to reflect the actual costs for a non U.S. purchase of 80 terminals by applying the same methodology as in section 6. That is, by using the 'fair' algorithm and a discount rate of 10% for both upper and lower limit RDE costs, and a slope factor of 92% for learning effects correction of the production costs for both upper and lower limits too. These are essentially the upper and lower limits of the cost ranges of uncertainty for

each procurement case. The total cost estimated range for the three alternative procurement routes is shown in Table 19 below.

Table 19. SYSTEM TOTAL COST ESTIMATED RANGE FOR 80 TERMINALS (\$M)

ALTERNATIVES	RANGE
Total Cost through FMS	(46.33, 70.96)
Total Cost through FMS and Prime Manufacturer	(42.07, 64.90)
Total Cost through Prime Manufacturer	(15.93, 28.36)

b. Sensitivity analysis for the discount rate

This part of the analysis aims to discover the degree of cost sensitivity with respect to the discount rate variations, over a range of assumed values. It must be noted that only RDE cost variations are analyzed because the historical production cost figures are already discounted. The results for 792 terminals are shown in Table 20.

Table 20. MEAN RDE COST VARIATIONS RELATIVE TO VARIOUS DISCOUNT RATES FOR 792 TERMINALS (\$M)

DIS-COUNT VALUES (r)	0	.03	.05	.10	.15	.18	.20
MEAN RDE COST	869.84	783.7	733.47	633.8	542.85	500.44	475.06

The figures then are converted to reflect the actual costs for a non U.S. purchase of eighty (80) terminals. Mean RDE cost, mean production cost and mean total cost values are reported in Table 21 for each procurement case. It must be noted that the third case of procurement, through the direct contact of prime manufacturer, corresponds to the mean production cost

Table 21. MEAN COST VARIATIONS RELATIVE TO VARIOUS DISCOUNT RATES FOR 80 TERMINALS (\$M)

MEAN RDE COST CASE1: FMS	50.7	45.6	42.7	36.95	31.65	29.18	27.7
MEAN PRO- DUCTION COST CASE1: FMS	22.08	22.08	22.08	22.08	22.08	22.08	22.08
MEAN TOTAL COST CASE1: FMS	72.78	67.68	64.78	59.03	53.73	51.26	49.78
MEAN RDE COST CASE2: FMS/PM	43.49	39.18	36.67	31.69	27.14	25.02	23.75
MEAN PRO- DUCTION COST CASE2: FMS/PM	22.08	22.08	22.08	22.08	22.08	22.08	22.08
MEAN TOTAL COST CASE2: FMS/PM	65.53	61.25	58.75	53.77	49.22	47.10	45.83
MEAN TOTAL COST CASE3: PM	22.08	22.08	22.08	22.08	22.08	22.08	22.08

value, because the RDE costs in this case are zero. This production cost value appears constant because it was already historically discounted.

c. Sensitivity analysis for the learning curve slope

The most typical slopes for electronic and communication systems are within the range from 80 to 99 percent [Ref. 23: p.37]. It must be noted that slope variations will only affect the production costs. The results of such an analysis are shown in Table 22.

Table 22. MEAN COST VARIATIONS WITH RESPECT TO VARIOUS SLOPE FACTORS FOR 80 TERMINALS (\$M)

SLOPE FACTORS	.80	.84	.88	.92	.96	.99
MEAN PRODUCTION COST	2.79	5.94	10.72	22.08	39.61	57
MEAN RDE COST CASE1: FMS	36.95	36.95	36.95	36.95	36.95	36.95
MEAN TOTAL COST CASE1: FMS	39.74	42.89	47.67	59.03	76.56	93.95
MEAN RDE COST CASE2: FMS/PM	31.69	31.69	31.69	31.69	31.69	31.69
MEAN TOTAL COST CASE2: FMS/PM	34.48	37.63	42.41	53.77	71.30	88.69
MEAN TOTAL COST CASE3: PM	2.79	5.94	10.72	22.08	39.61	57

d. Combined sensitivity analysis

The effect on total cost behavior as a consequence of the RDE cost variations with respect to several discount rates and a fixed slope factor of 92 percent, for each of the three alternative routes of procurement, is shown in Figures 73, 74, 75. The respective effect due to the slope factor variations and a fixed discount rate of 10 percent is also shown in the same figures for the same three cases of procurement. Finally, these figures show the effect of cost uncertainty. In a sense, they present the overlapping effect on the total cost of all kinds of parameter variations. This composite effect is essentially represented pictorially by an overall area of uncertainty.

8. Results

The following conclusions are drawn from this LCC analysis.

- A system procurement based on the JTIDS technology and design will be most reasonable if there is a possibility to directly contact the prime manufacturers, provided that such a route is feasible. The feasibility of such an alternative depends on the degree of existing competition in the open market and the inherent barriers of technology transfer. This procurement procedure would minimize the R&D-to-investment ratio.
- A design based on the JTIDS-concept and technology may exceed the requirements necessary for smaller scale systems to be used in different environments. This might encourage the decision maker to set specifications for a totally original design, with less capacity transmitting power, hardware interfaces and nuclear hardening of the terminals used. Although production costs may normally decrease exponentially with these parameters, this option is regarded as not affordable for the time being. Because there isn't yet any competition in the open market for such a system with fully integrated C3 anti-jam tactical communication capabilities, the R&D costs would by far exceed the most unfavorable JTIDS figures. In fact JTIDS is the first of its kind.
- Procurement of such a system late in the future is consistent with the technological capabilities of the potential threat, and the prospects of an open competitive market. It also follows the mathematical realities for a cheaper design since the production costs normally drop and the discounting process is cumulative. Furthermore, original R&D costs may have already been fully allocated. In joint procurements such as JTIDS, the later the customer joins the project the lower are the R&D costs allocated to him. But there is a trade-off in delaying the purchase. There are opportunity costs that customers incur due to the delay in acquiring the new technology and taking advantage of the relevant benefits. The decision maker must evaluate the costs and benefits before he makes his final decision.
- A life cycle cost analysis is only objective and realistic if the figures used are objective and realistic. The projection of communication systems costs is heavily based upon historical data from previous similar engineering efforts. This is particularly relevant in this case where the system is a first-of-a-kind system. The decision maker must carefully examine and analyze the accuracy of such figures. Moreover, he must justify the discount rates and the learning curve slope factors. Results can be easily skewed when arbitrary slope factors, subjective discount rates and misleading historical data is used.
- It is up to the decision makers to choose one of the possible three alternative cases for this system's procurement. This decision, of course, is dependent on all the situational assumptions mentioned before. Whatever decision is made, it will only be informed if these situational factors and

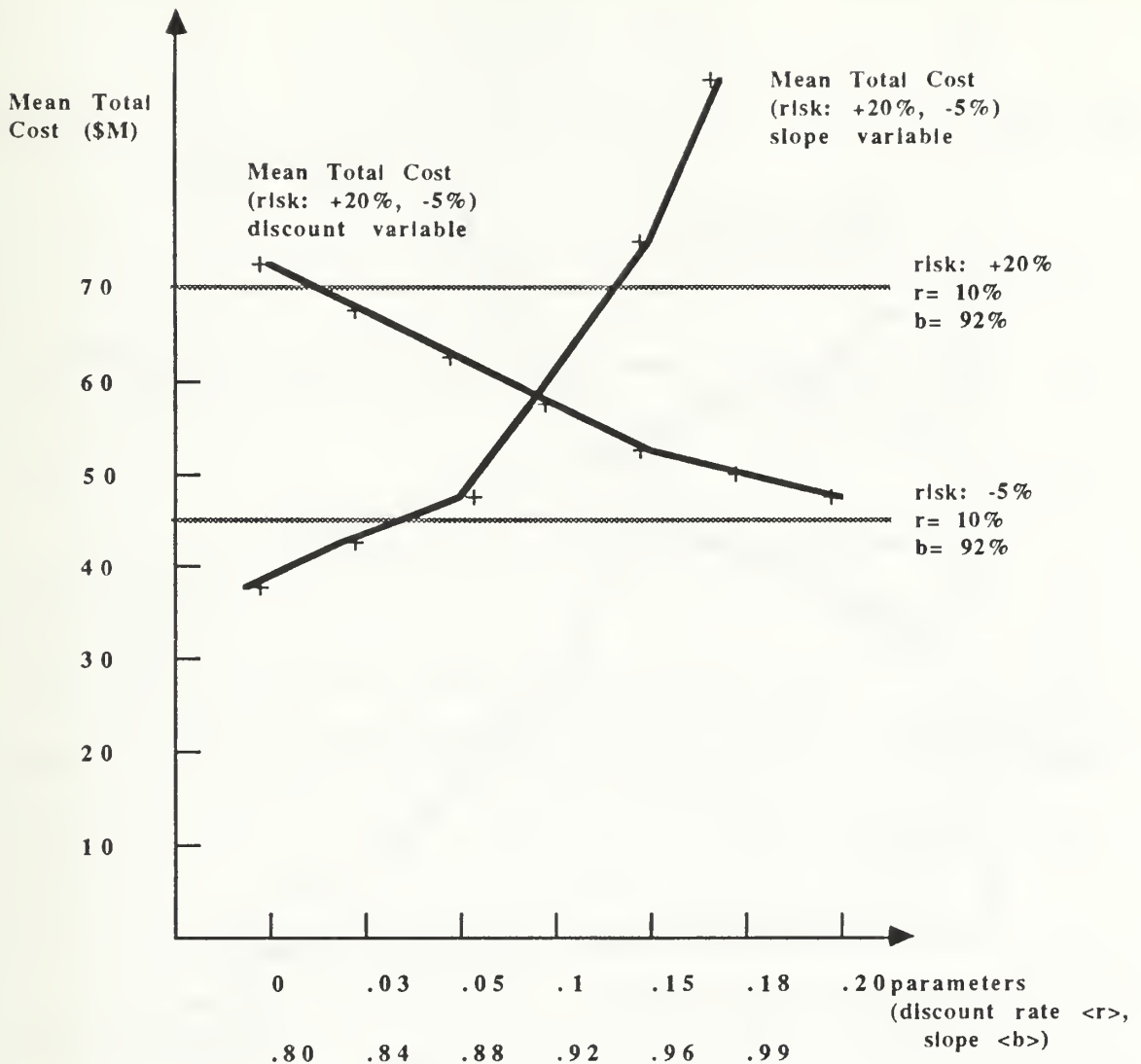


Figure 73. Total Cost Behavior (Case 1: FMS): overall area of uncertainty.

assumptions are carefully weighed. After weighing these factors, the choice should lie with that alternative which provides an acceptable cost with the highest probability. Although the most likely cost may be higher than the cost of another alternative, it may still be preferred if the lower cost has a lower probability or a larger area of uncertainty.

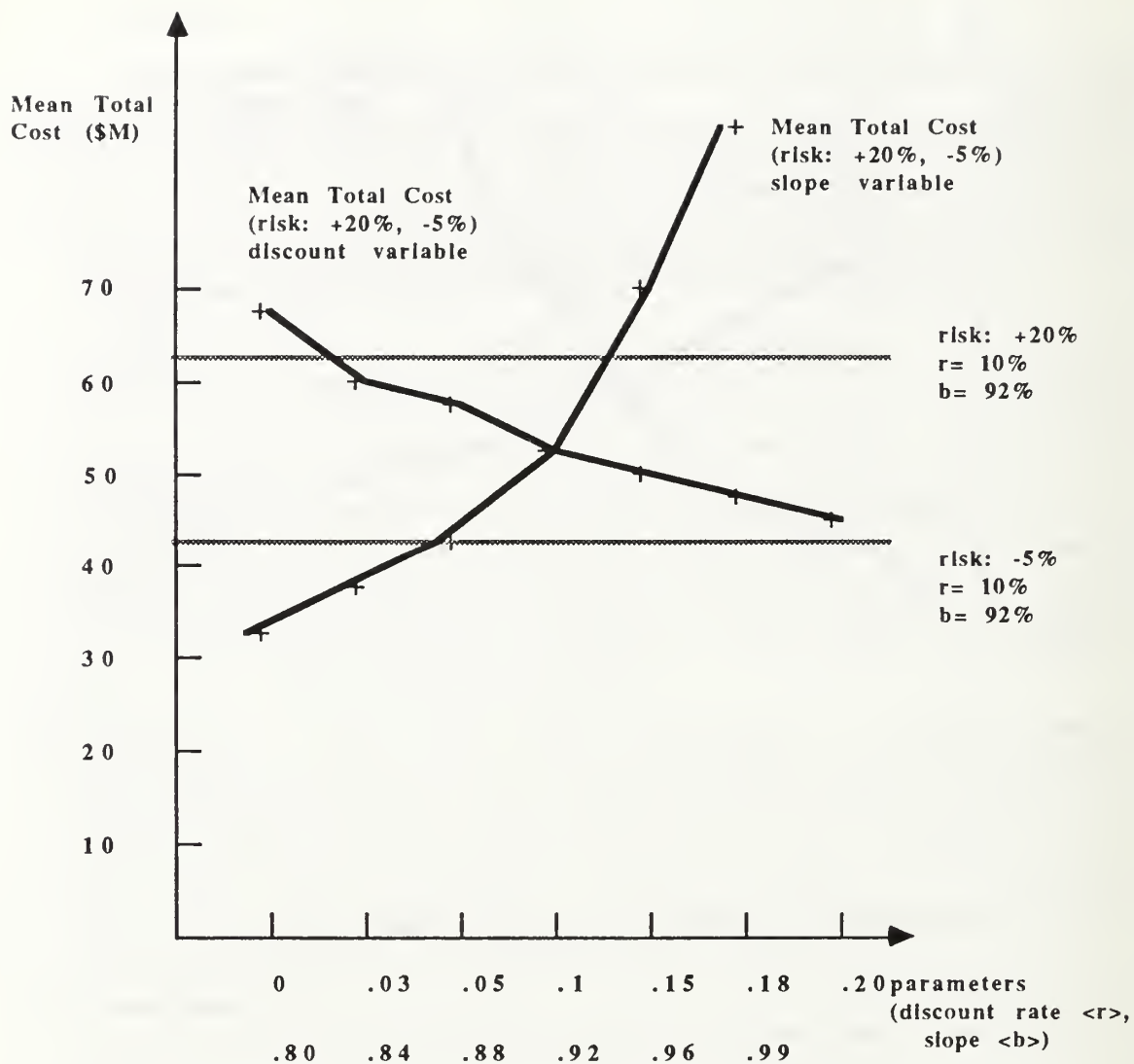


Figure 74. Total Cost Behavior (Case 2: FMS and PM): overall area of uncertainty.

Mean Total
Cost (\$M)

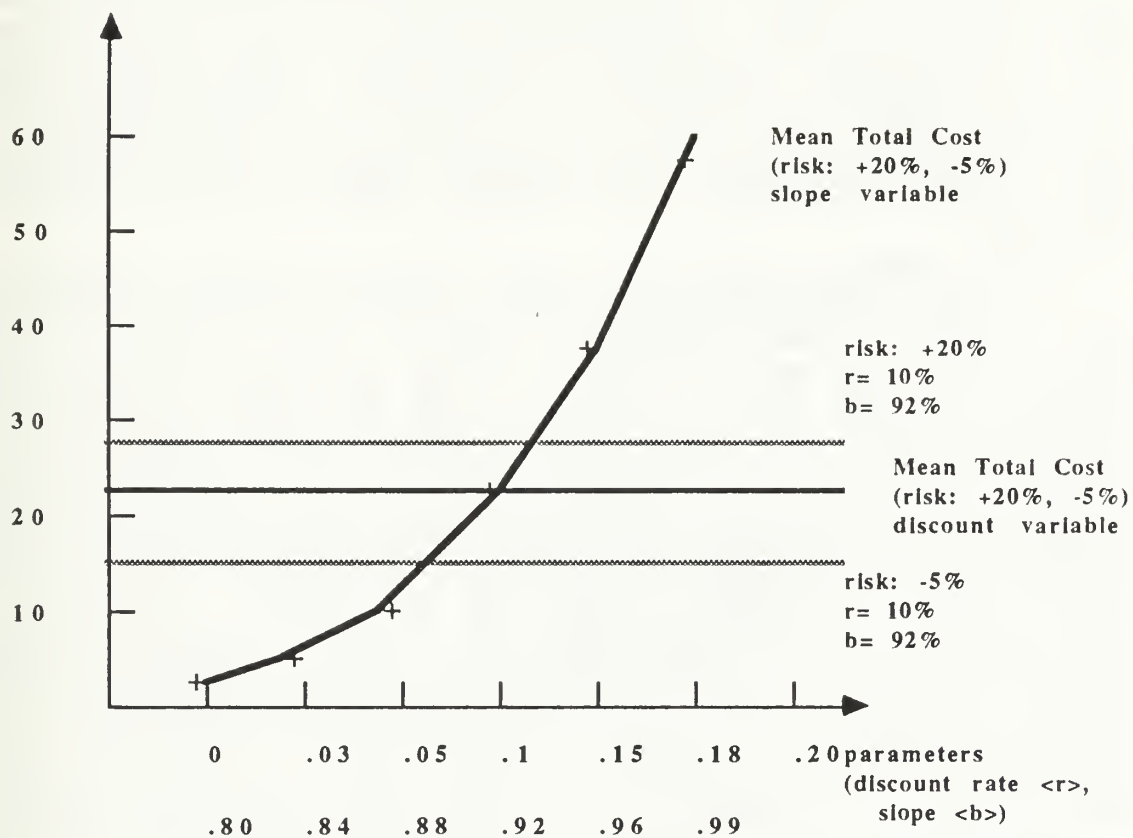


Figure 75. Total Cost Behavior (Case 3: PM): Overall area of uncertainty.

VI. CONCLUSIONS AND RECOMMENDATIONS

A. CONCLUSIONS

Effective coordination of all forces involved in joint tactical naval operations in closed-sea areas can be achieved through real-time, survivable, secure and high capacity communications. A joint tactical digital communication link can fulfill these requirements and it can provide interoperable solutions to assist the Naval Commander in the command, control and communications (C3) process.

A conceptual design based on the Joint Tactical Information Distribution System (JTIDS) provides real-time, high capacity (up to 300 kbps), multi-function (up to 128), multi-net (up to 16) voice and data communications through the use of time division/distributed time division multiple access (TDMA/DTDMA) technology. Functions such as Link 11, Link 4A, Relative Navigation, Precise Positioning (up to 100 ft), IFF (identification of friendly and foe) and tactical air navigation (TACAN) are also provided. TDMA transmission takes place during fixed assigned time slots. By permitting variable time slots, DTDMA improves the system efficiency, since every potential station uses only the necessary portion of the total system capacity. Both modes allow many users simultaneous access to the link. System effective throughput is limited by the use of forward error correcting and detecting schemes.

The system gracefully degrades due to its non-nodal connectivity nature; consequently, its survivability is enhanced to a degree that communications are maintained in the event of several system platform losses. It is also very resistant to enemy jamming and electronic support measures (ESM) activity due to the use of spread spectrum technology. Although costly, this technology is very effective against enemy electronic counter-measures (ECM) and ESM activity, under current realities, as it provides high anti-jam system margins and low probability of intercept (LPI) levels. Furthermore, the system is very secure due to the employment of advanced message coding schemes.

Jamming analysis shows that the jammer has to operate from practically unrealistic distances in order to jam the system effectively, when barrage jamming is intended. Geometrical situations are nearly always such that the jammer is at a severe disadvantage compared to a desired transmitter as far as the path attenuation factors are concerned. Increased jamming power could overcome the inherent system high AJ margin and the path attenuation, but it might also affect the jamming platform's flexibility. Theoretical results show that barrage jamming by an airborne or a shipborne conventional jammer against this system is not possible from realistic distances. Although the need for the use of omnidirectional antennas by the system makes it more susceptible to highly directional jamming, this is not a serious problem. The lack of antenna null-forming capability is offset by the high system processing gain. Airborne jamming has an advantage over shipborne jamming due to the relative unpredictability of the jammer. However, it is not effective because of platform power-carrying limitations, the high system AJ margin and the combined attenuation factors encountered along the jamming path. In fact, the jammer will always be within the system's receiver air-to-surface missile (ASM) range due to the inherent high processing gains provided by the hybrid spread spectrum modulation and the coding techniques employed by the system. Furthermore, shipborne barrage jamming is not possible against this system because the jammer platform will also be within the system's platform surface-to-surface missile (SSM) range, even if higher jamming power can be put out by this kind of jammer.

Smart jamming is another option in attempting to jam this system effectively but it requires fast and smart repeaters with high jamming power output. An effective alternative might be to jam the synchronization circuit of the system. This option requires prior knowledge of certain system parameters before any action is taken by the potential jammer. In any case, the cost-effectiveness of building a smart jammer for this system is certainly questionable and it is probably an area of further research. On the other hand, a JTIDS-based design should not be viewed as an unjammable, cure-all sys-

tem for all the problems that may be encountered. However, the jammer can be presented with such problems that its task becomes practically difficult and expensive.

System link analysis calculations show that air-to-ground communication ranges are satisfactory under realistic environmental conditions (up to 300 NM). Surface-to-surface ranges are essentially line-of-sight (up to 20 NM) for a UHF system like this. This is a major flaw for the system because the longer command, control and communications (C3) paths normally required for joint tactical operations should be served either by relays or by using existing HF data links (Link 11, Link 4A, Link 14). But these links have inherent disadvantages such as jamming susceptibility, slow data rates and dual tracking errors, and the use of relays does not appear to be a very attractive solution. This is because continuous availability of relays is not feasible for smaller scale navies even if certain solutions such as the use of conventional maritime patrol aircraft (MPAs) are employed. The conventional MPA, with its limited effective operating time and range, presents availability problems even if the modification cost and engineering effort for the respective terminal are not an issue.

The choice of the Lx portion of the UHF band for system operation is based on its relatively low utilization, the favorable air-to-ground line-of-sight characteristics, and the compatibility with existing hardware serving other functions such as IFF and TACAN. Given these advantages, and the range disadvantages mentioned before, it is believed that future C3 systems with comparable technology and beyond line-of-sight ranges (without relays) must be an area of potential research.

Deployment analysis and planning should consider ways to identify the most efficient system configuration. Configuration depends on certain system parameters, such as power settings, antennas, and the impact that the environmental and operating terrain element has on the attenuation loss to be used. Maximum achievable ranges with the worst case conditions can then be calculated and used in the decision making process by the Naval Commander. Spatial deployment separation must be considered as well, due to the "near-

far” effect. Closely spaced units can operate with high power output when the employment of hybrid spread spectrum techniques is based on accurate synchronization. Also, code division multiple access (CDMA) guarantees the avoidance of mutual interference of closely spaced system-equipped units when perfectly uncorrelated PN coding schemes are employed.

Affordability for such a system is questionable for smaller navies due to the lack of competition in the related area, for the time being. Although the capacity, the power and EMP hardening levels of a JTIDS-based design by far exceeds the respective levels required by smaller scale navies, the high R&D-to-Investment ratio renders an original design uneconomic for the time being. The procurement of a totally new system in the future is consistent with the technological capabilities of the threat and the lower production costs due to the learning effects. With all options present, procurement of such a system directly through the prime manufacturer appears to be the cheapest route, because it minimizes the associated R&D-to-Investment ratio. Acquisition of a JTIDS-based system through official routes appears a preferable solution for the long-term future. R&D cost allocations for joint projects follow a declining pattern, being proportional to the number of prior potential system customers. In any case, the decision maker should evaluate the trade-offs between a purchase delay and a timely acquisition of the new technology in such a way that opportunity costs won't be incurred. Furthermore, when a life cycle cost analysis is performed, the discount rate and learning curve slope factors, as well as the historical cost figures, must be carefully examined and justified using sensitivity analysis. Otherwise, a false decision might be reached.

JTIDS is not still fully operational and it has a development window of about 15 years. This may lead to a situation where the technology employed by the system will no longer be state-of-the-art after it is fully tested and deployed. A 10-year-old anti-jam (AJ) margin performance requirement may be meaningless as the threat develops new ECM capabilities.

B. RECOMMENDATIONS

Based on the overall analysis attempted in this study the following are recommended as areas for further research:

- The conceptual design of a repeater jammer capable of attacking a highly (AJ) system like the proposed one.
- The conceptual design of a 'smart' jammer capable of attacking the synchronization circuit of the system.
- And, the examination of the design feasibility of an HF AJ system capable of providing long-range high capacity communications as compared to the line-of-sight ranges achievable by the proposed system.

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